VOLUME 107 PART B NUMBER 31 JANUARY 1960

J. T. LUDWIG
APR 23 1960
SCIENTIFIC ADVISORY GROUP



The Proceedings

OF

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871: INCORPORATED BY ROYAL CHARTER 1921

PART B

ELECTRONIC AND COMMUNICATION ENGINEERING
(INCLUDING RADIO ENGINEERING)

SAVOY PLACE · LONDON W.C.2

Price Ten Shillings and Sixpence

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871 INCORPORATED BY ROYAL CHARTER 1921

PATRON: HER MAJESTY THE QUEEN

COUNCIL 1959–1960

President

SIR WILLIS JACKSON, D.Sc., F.R.S.

Past-Presidents

W. H. ECCLES, D.Sc., F.R.S.
THE RT. HON. THE EARL OF MOUNT EDGCUMBE, T.D.
J. M. DONALDSON, M.C.
PROF. E. W. MARCHANT, D.Sc.
H. T. YOUNG.
SIR GEORGE LEE, O.B.E., M.C.
SIR ARTHUR P. M. FLEMING, C.B.E., D.Eng., LL.D.
J. R. BEARD, C.B.E., M.Sc.
SIR NOEL ASHBRIDGE, B.Sc.(Eng.).
SIR HARRY RAILING, D.Eng.
P. DUNSHEATH, C.B.E., M.A., D.Sc. (Eng.), LL.D.
SIR VINCENT Z. DE FERRANTI, M.C.
T. G. N. HALDANE, M.A.
PROF. E. B. MOULLIN, M.A., Sc.D., LL.D.
SIR ARCHIBALD J. GILL, B.Sc.(Eng.).
SIR JOHN HACKING.
COL. B. H. LEESON, C.B.E., T.D.
SIR HAROLD BISHOP, C.B.E., B.Sc.(Eng.), F.C.G.I.
SIR JOSIAH ECCLES, C.B.E., D.Sc.
SIR GEORGE H. NELSON, Bart.
SIR GORDON RADLEY, K.C.B., C.B.E., Ph.D.(Eng.).
S. E. GOODALL, M.Sc.(Eng.), F.Q.M.C.

Vice-Presidents

O. W. HUMPHREYS, C.B.E., B.Sc. G. S. C. LUCAS, O.B.E., F.C.G.I. SIR HAMISH D. MACLAREN, K.B.E., C.B., D.F.C., LL.D., B.Sc. C. T. MELLING, C.B.E., M.Sc.Tech. A. H. MUMFORD, O.B.E., B.Sc.(Eng.).

Honorary Treasurer

E. LEETE.

Ordinary Members of Council

PROF. H. E. M. BARLOW, Ph.D., B.Sc. (Eng.).
C. O. BOYSE, B.Sc.(Eng.).
PROF. M. W. HUMPHREY DAVIES, M.Sc.
SIR JOHN DEAN, B.Sc.
L. DRUCQUER.
J. M. FERGUSON, B.Sc.(Eng.).
D. C. FLACK, B.Sc.(Eng.), Ph.D.
J. S. FORREST, D.Sc., M.A.
R. J. HALSEY, C.M.G., B.Sc.(Eng.), F.C.G.I.
J. B. HIGHAM, Ph.D., B.Sc.
R. A. HORE, M.A., B.Sc.
F. C. MCLEAN, C.B.E., B.Sc.
B. L. METCALF, B.Sc.(Eng.).
J. R. MORTLOCK, Ph.D., B.Sc.(Eng.).
H. G. NELSON, M.A.
R. H. PHILLIPS, T.D.
H. V. PUGH.
J. R. RYLANDS, M.Sc., J.P.
G. A. V. SOWTER, Ph.D., B.Sc.(Eng.).
C. E. STRONG, O.B.E., B.A., B.A.I.
D. H. TOMPSETT, B.Sc.(Eng.).

Chairmen and Past-Chairmen of Sections

Electronics and Communications:
M. J. L. PULLING, C.B.E., M.A.
*G. MILLINGTON, M.A., B.Sc.

Measurement and Control:
PROF. A. TUSTIN, M.Sc.
*J. K. WEBB, M.Sc.(Eng.), B.Sc.Tech.

Supply:
J. R. MORTLOCK, Ph.D., B.Sc.(Eng.).
*D. P. SAYERS, B.Sc.

Utilization:
T. E. HOUGHTON, M.Ent.
*R. A. MARRYAT, B.Sc.(Eng.).

Chairmen and Past-Chairmen of Local Centres

East Midland Centre:
D. H. PARRY, B.Sc.
*D. E. LAMBERT, B.Sc.(Eng.).

North-Western Centre:
F. H. HUTCHINSON, M.Eng.
*PROF. F. C. WILLIAMS, O.B.E., D.Sc., D.Phil., F.R.S.

Mersey and North Wales Centre:
T. A. P. COLLEDGE, B.Sc.(Eng.).
*J. COLLINS.

Northern Ireland Centre:
T. S. WYLIE.
*D. S. MCILHAGGER, Ph.D., M.Sc.

Southern Centre:
W. D. MALLINSON, B.Sc.(Eng.).
*G. BISHOP, B.Sc.

North-Eastern Centre:
H. WATSON-JONES, M.Eng.
*A. T. CRAWFORD, B.Sc.

Scottish Centre:
J. A. AKED, M.B.E.
*R. J. RENNIE, B.Sc.

Western Centre:
H. JACKSON, B.Sc.(Eng.).
*R. W. STEEL.

North Midland Centre:
PROF. G. W. CARTER, M.A.
*J. D. NICHOLSON, B.Sc.

South Midland Centre:
G. F. PEIRSON.
*L. L. TOLLEY, B.Sc.(Eng.).

* Past-Chairman

ELECTRONICS AND COMMUNICATIONS SECTION COMMITTEE 1959–1960

Chairman

M. J. L. PULLING, C.B.E., M.A.

Vice-Chairmen

R. J. HALSEY, C.M.G., B.Sc.(Eng.), F.C.G.I. T. B. D. TERRONI, B.Sc. J. A. RATCLIFFE, O.B.E., M.A., F.R.S.

Past-Chairmen

G. MILLINGTON, M.A., B.Sc. J. S. MCPETRIE, Ph.D., D.Sc.

Ordinary Members of Committee

D. A. BARRON, M.Sc.
P. A. T. BEVAN, C.B.E., B.Sc.
J. BROWN, M.A., Ph.D.
PROFESSOR A. L. CULLEN, Ph.D., B.Sc.(Eng.).
L. I. FARREN, M.B.E.
G. G. GOURIET.
COMDR. C. G. MAYER, O.B.E., U.S.N.R.
J. MOIR
L. J. I. NICKELS, B.Sc.(Eng.).
B. G. PRESSEY, M.Sc.(Eng.). Ph.D.
N. C. ROLFE, B.Sc.(Eng.).
T. R. SCOTT, D.F.C., B.Sc.
C. WILLIAMS, B.Sc.(Eng.).
W. E. WILLSHAW, M.B.E., M.Sc.Tech.
A. J. YOUNG, B.Sc.(Eng.).

And

The President (*ex officio*)
The Chairman of the Papers Committee.
PROFESSOR H. E. M. BARLOW, Ph.D., B.Sc.(Eng.) (representing the Council).
E. H. COOKE-YARBOROUGH, M.A. (Co-opted Member).
G. H. HICKLING, B.Sc. (representing the North-Eastern Measurement and Electronics Group).
K. F. SANDER, M.A., Ph.D., B.Sc. (representing the Cambridge Electronics and Communications Group).
J. STEWART, M.A., B.Sc. (representing the Scottish Electronics and Measurement Group).
R. FEINBERG, Dr.Ing., M.Sc (representing the North-Western Electronics and Communications Group).
R. E. YOUNG, B.Sc.(Eng.) (representing the South Midland Electronics and Measurement Group).

The following nominees:
Royal Navy: CAPTAIN W. D. F. B. MUSPRATT, R.N.
Army: COL. R. G. MILLER, M.A.
Royal Air Force: GROUP CAPTAIN D. W. ROWSON, B.Sc.(Eng.), R.A.F.

MEASUREMENT AND CONTROL SECTION COMMITTEE 1959–1960

Chairman

PROFESSOR A. TUSTIN, M.Sc.

Vice-Chairmen

C. G. GARTON; W. S. ELLIOTT, M.A.

Past-Chairmen

J. K. WEBB, M.Sc.(Eng.), B.Sc.Tech.; H. S. PETCH, B.Sc.(Eng.).

Ordinary Members of Committee

E. W. CONNON, B.Sc.(Eng.), M.Eng.
A. C. LYNCH, M.A., B.Sc.
A. J. MADDOCK, D.Sc.
R. E. MARTIN.
A. NEMET, Dr.Sc.Techn.
S. N. POCOCK.
W. RENWICK, M.A., B.Sc.
G. A. V. SOWTER, Ph.D., B.Sc.(Eng.).
G. F. TAGG, Ph.D., B.Sc.
R. D. TROTTER, B.Sc.(Eng.).
J. H. WESTCOTT, B.Sc.(Eng.), Ph.D.
F. C. WIDDIS, B.Sc.(Eng.).

And

The President (*ex officio*).
The Chairman of the Papers Committee.
G. A. V. SOWTER, Ph.D., B.Sc.(Eng.) (representing the Council).
C. C. BAXENDALE (representing the North-Eastern Measurement and Electronics Group).
E. J. R. HARDY, B.Sc.(Eng.) (representing the North-Western Measurement and Control Group).
D. L. A. BARBER, B.Sc.(Eng.) (nominated by the National Physical Laboratory).
H. M. GALE, B.Sc.(Eng.) (representing the South Midland Electronics and Measurement Group).
W. H. P. LESLIE, B.Sc. (representing the Scottish Electronics and Measurement Group).

Secretary

W. K. BRASHER, C.B.E., M.A., M.I.E.E.

Deputy Secretary

F. JERVIS SMITH, M.I.E.E.

Principal Assistant Secretary

F. C. HARRIS.

Editor-in-Chief

G. E. WILLIAMS, B.Sc.(Eng.), M.I.E.E

THE WASHING STAGE IN SEMI-CONDUCTOR MANUFACTURE

Issued by *ELGA PRODUCTS LIMITED*

The analysis below represents intrinsic* water, considered essential for rinsing transistors and similar devices. There are three methods of producing this very high purity water. Whatever method is used, effluent quality must conform to the analysis.

* intrinsic: belonging naturally—inherent—essential

METHOD 1

Individual line stations—mixed bed column

Observations: Effluent is conductivity water approaching 10 megohm-cm, no regeneration *in situ*. Appears to be satisfactory for most applications.

Limitation: Rather uneconomical in hard water areas.

METHOD 2

***Central plant with regeneration* in situ, connected to end purification stations**

Observations: Effluent quality is "intrinsic water" of the order of 16 megohm-cm+. This approach appears to be generally acceptable and is very economical.

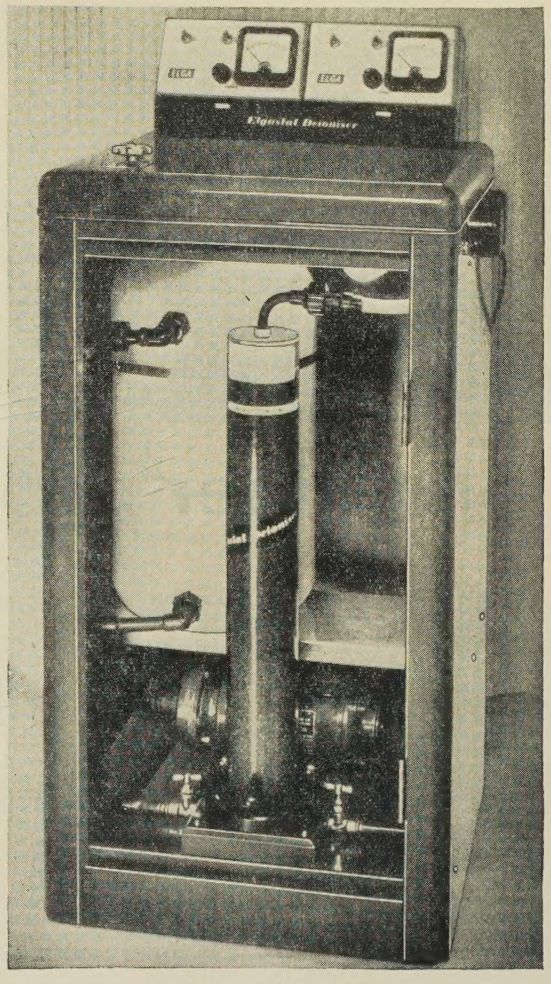
METHOD 3

Individual line stations—mixed bed recirculation

This method produces intrinsic water of the order of 16 megohm-cm+. No regeneration *in situ* or central plant required.

Observations: Acceptable effluent for all known applications, virtually no maintenance. Negligible effluent cost: £2. 10. 0. for 6 to 10 working days. An example of a mixed bed recirculator is shown. This model cost £187.

The Company specialises in the design and manufacture of deionisers for the provision of intrinsic water. Publication IEE/2 includes a paper "Intrinsic Water for Semi-conductor Washing". Please ask for a copy.



The Elgastat Robot B.106.

DEIONISED EFFLUENT ANALYSIS	RESULTS
Trace metals	*Not present*
Silica	*Below 0.1 p.p.m.*
Chloride	*Not present*
Sulphate	*Not present*
Ammonia	*No coloration with Nessler's reagent*
Carbon dioxide	*Not present*
pH value	*6.6-7*
Electrical resistance	*10 megohm-cm+*

If it's ion exchange ask ELGA first

ELGA PRODUCTS LTD

LANE END BUCKS Telephone: Lane End 396

a

we can get you out of a rectifier problem

We, that's the M-O Valve Co. Ltd., have spent over 40 years bottling up hard valves, soft valves, tough going-aloft valves, standard valves, valves to government specs . . . so we know how you feel in there.

But we're used to it, because now we've acquired a rare collective know-how that can beat most valve problems. And of course, it's backed by the brilliant research, progressive development and flexible production you'd expect us to have.

Next time you're enveloped in a valve problem, let us know—if we haven't a ready-made type for you, we'll see if we can make you one to measure.



XENON FILLED TYPE. *The G.E.C. range will keep going a long time—and has better PIV safety ratings too.*

MERCURY VAPOUR TYPE. *We've had over 30 years of know-how in making this type of rectifier. As you'd expect, they're the best you can get. There's a range of eight.*

VACUUM TYPE. *Too many even to give you an outline. (Bi-phase, full-wave, high voltage, special quality, 450 to 28,000V rms.)*

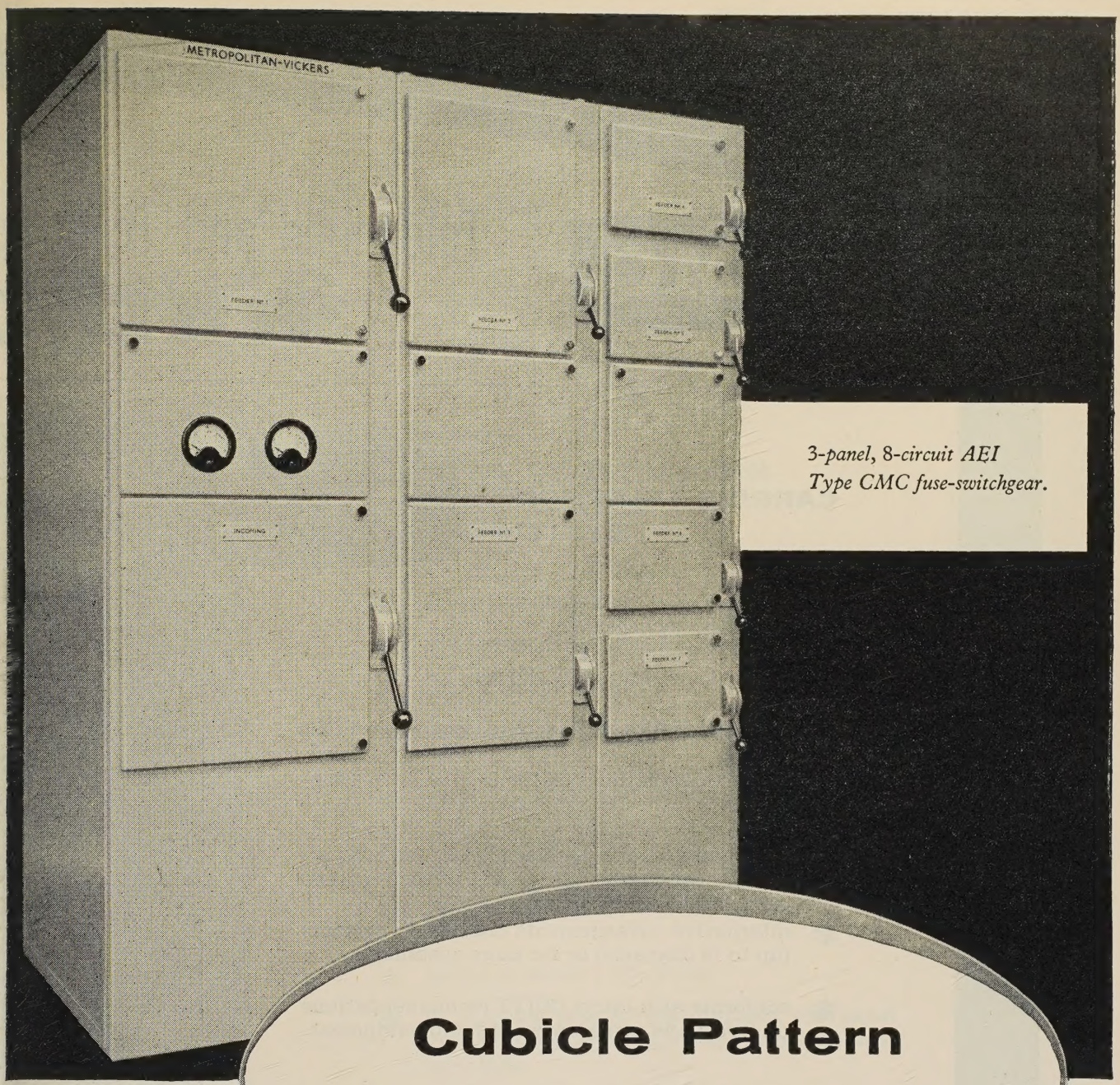
Let us know what types you're interested in and we'll send you full details.

G.E.C. valves are obtainable from

THE M-O VALVE CO. LTD

BROOK GREEN · HAMMERSMITH · LONDON W.6.

A subsidiary of the General Electric Co. Ltd



3-panel, 8-circuit AEI Type CMC fuse-switchgear.

Cubicle Pattern Fuse-Switchboards

Type CMC fuse-switchgear is designed to meet modern requirements of totally enclosed low-voltage distribution switchboards.
Installation is quick and simple and additional circuits are easily provided by connecting further cubicle sections.
H.R.C. cartridge-type fuses are fitted throughout.
For full details write for Publication 299/5-1

AEI

Made at the Trafford Park Works of

Associated Electrical Industries Limited
SWITCHGEAR DIVISION
Trafford Park, Manchester · Higher Openshaw, Manchester · Willesden, London

F/A906

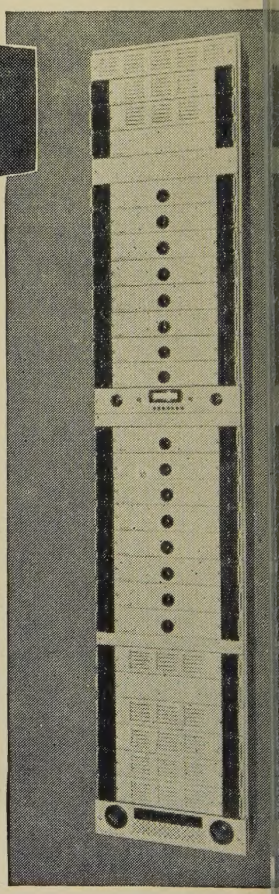
TMC

Telephone and Telegrap

RADIO

CARRIER TELEPHONE SYSTEM

R40C



- 6 kc/s channel spacing
- channel bandwidth 300-3400 c/s
- out-band dialling (E & M) or ringdown facilities
- low-level signalling (-20dbm0) tone-on or tone-off idle
- 40 circuits on one double-sided 9 foot rack including carrier, signalling and power supplies
- alternative arrangements for small systems (up to 16 channels) or for large systems
- conforms with latest CCITT recommendations on levels and impedance for radio equipment

THESE TWO SYSTEMS ARE PART OF THE **T.M.C.** RANGE OF TRANSMISSION EQUIPMENT WHICH INCLUDES:

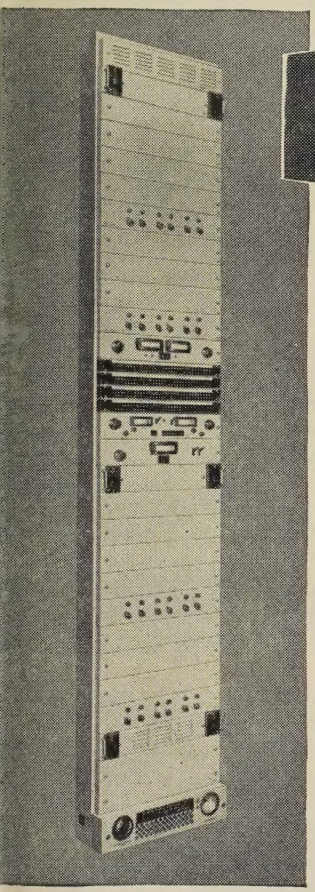
2, 3, 4 and 6 kc/s spaced carrier telephone equipment for cable, radio and ope wire systems. Transistorized 120 c/s and 170 c/s FM Telegraph Equipment. Transistori VF Repeater Equipment. Transistorized privacy equipment.

TELEPHONE MANUFACTURI

TRANSMISSION DIVISION · CRAY WO

channelling equipment for...

LINKS



T24D **VF TELEGRAPH SYSTEM**

- Fully transistorized
- Frequency modulated (±30c/s)
- 120 c/s spaced channels
- 24 bothway channels on one speech circuit (300-3,400 c/s)
- Pilot frequency control, if required
- Plug-in channel units simplify maintenance and extensions
- Robust circuits using only 7 transistors per channel
- Mains or battery operation
- 24 channels with pilot equipment, test panel and double current telegraph supplies on 9′ rack-side

ELLING AGENTS

USTRALIA AND NEW ZEALAND — Telephone Manufacturing Co. (A'sia) Pty. Ltd. Sydney, N.S.W.

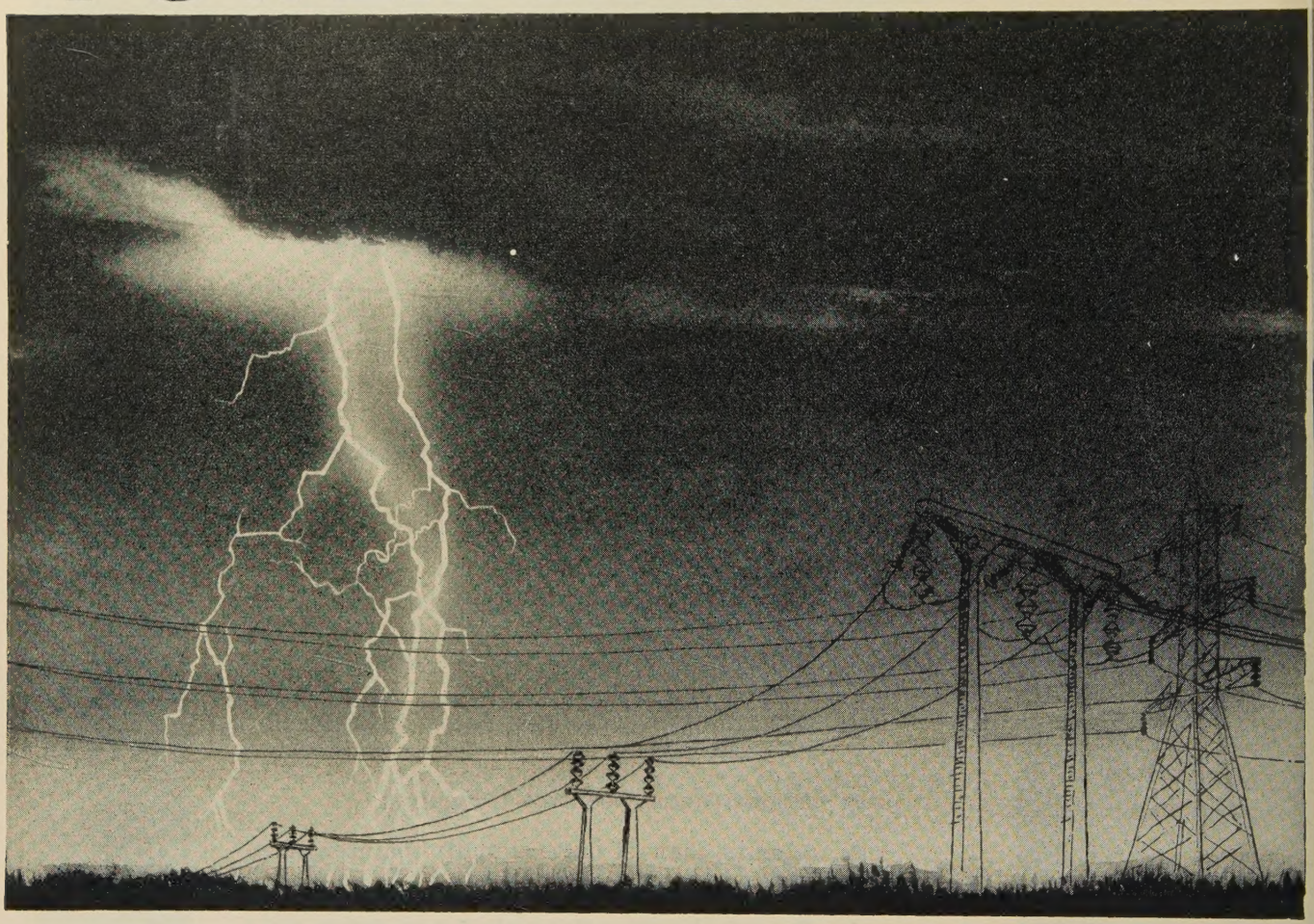
ANADA AND U.S.A. — Telephone Manufacturing Co. Ltd., Toronto, Ont.

LL OTHER COUNTRIES *(For Transmission Equipment Only)* — Automatic Telephone & Electric Co. Ltd., London.

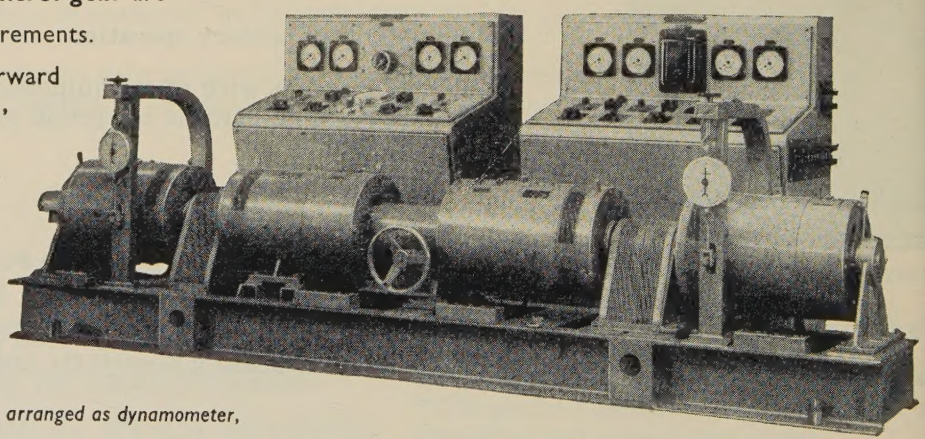
COMPANY LIMITED

EVENOAKS WAY · ORPINGTON · KENT

POWER AND CONTROL



E.D.C.C. machines and control gear are available to meet most requirements. Standard units for straightforward applications and 'tailor made' specials for the most diverse uses.



Recently supplied to the Welsh College of Advanced Technology, Cardiff—a multiple motor generator set, comprising 4 machines and two control desks. The machine unit comprises:—

- *7 h.p. D.C. variable speed motor, arranged as dynamometer, complete with Tacho-generator*
- *4kVA alternator*
- *4kVA alternator with rotable stator*
- *7 h.p. synchronous motor (left) arranged as dynamometer*
- *The complete unit can be split into two motor-alternator sets*

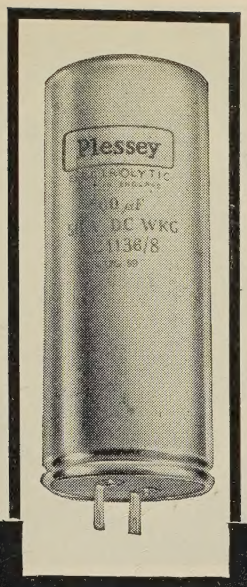
ELECTRO-DYNAMIC

CONSTRUCTION COMPANY LIMITED

ST. MARY CRAY • ORPINGTON • KENT • TELEPHONE: ORPINGTON 27551 • GRAMS: ELEDAMIC ST. MARY CRAY

CONTROL GEAR DIVISION: BRIDGWATER SOMERSET TEL: BRIDGWATER 2882 GLASGOW OFFICE: 40 HOULDSWORTH ST C3 TEL: CENTRAL 2620

P.2827



Plessey

HYPERLYTICS*
give
ten years service

Long shelf life and a working life of 10 years are achieved in the new type of electrolytic capacitor developed by Plessey.

The exacting specifications demanded for specialised electronic equipment have been met by using a high purity etched foil material, and combining the direct Plessey process control with unique manufacturing methods.

With low leakage and high insulation characteristics, Hyperlytics can be supplied in a wide range of values between 0.5 and 12,000 μF. Two types are available, the 1000 series with a temperature range of —30°C to +85°C, and the 11,000 series with a range of —20°C to +50°C.

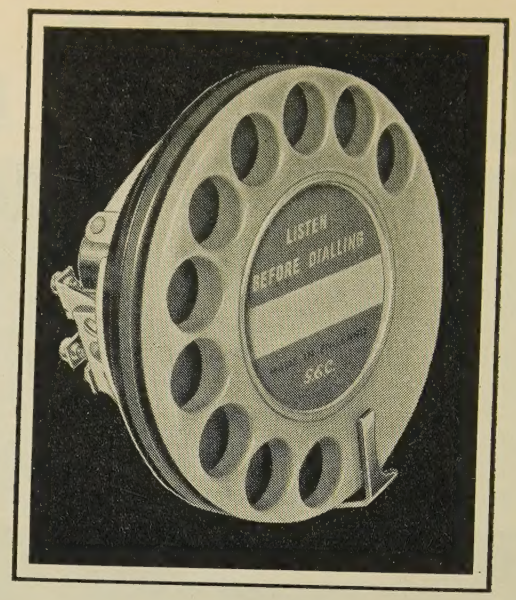
Capacitance tolerance for both types is normally —20% to +50% but closer limits can be achieved to meet special requirements.

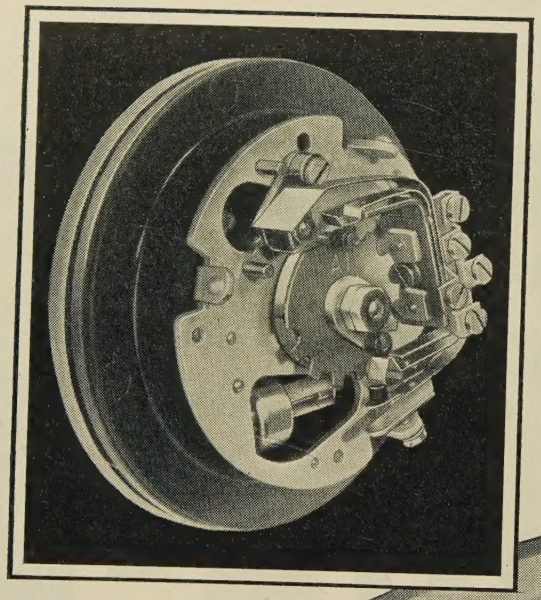
Write for further details of these outstanding new components.

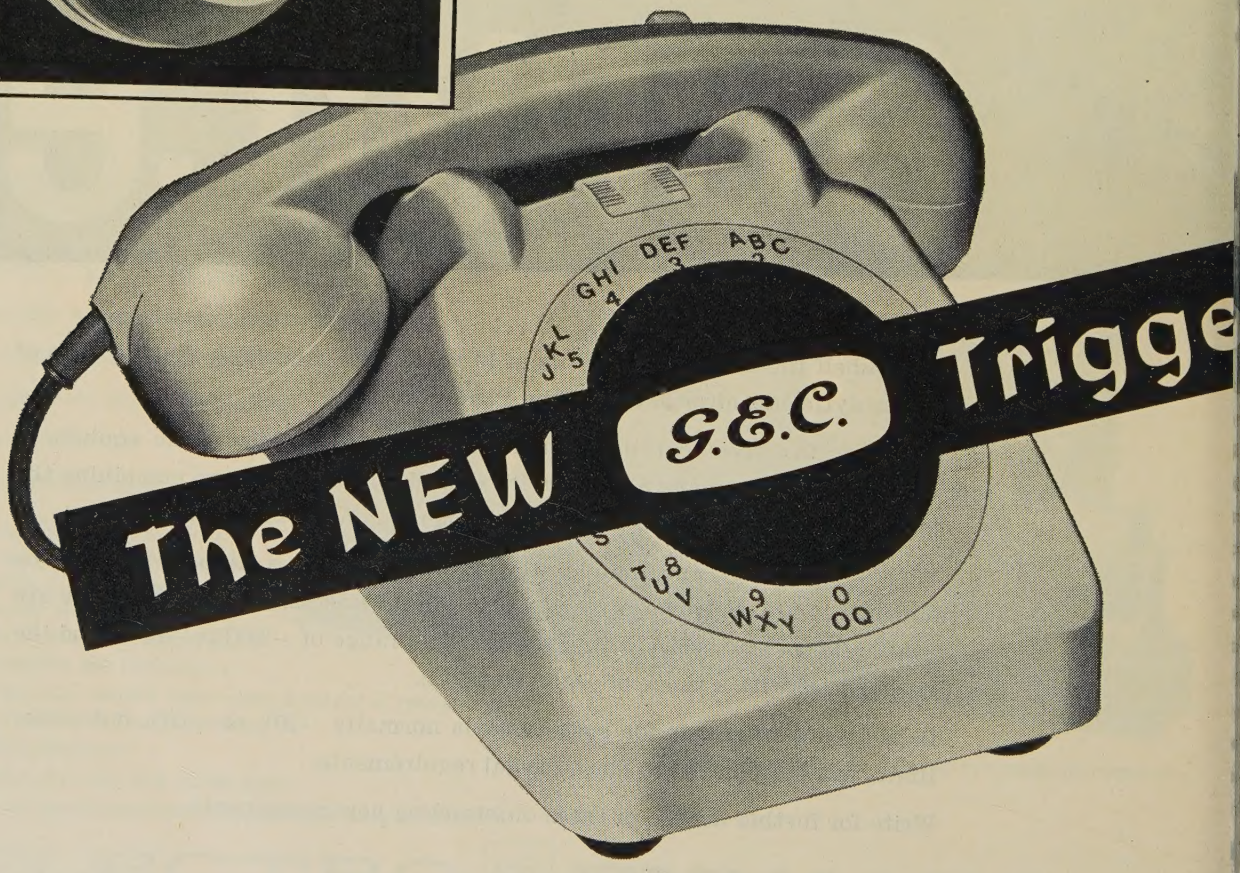
* *High Performance Electrolytic Capacitors*

THE PLESSEY COMPANY LIMITED · CAPACITORS & RESISTORS DIVISION
KEMBREY STREET · SWINDON · WILTS · TELEPHONE: SWINDON 6211
Overseas Sales Organisation: Plessey International Limited · Ilford · Essex · Telephone: Ilford 3040

CRI/A


LISTEN
BEFORE DIALLING
G.E.C.




The NEW G.E.C. Trigge
GHI DEF ABC
JKL
TUV
WXY
OQ

NDARDISED BY THE BRITISH POST OFFICE
PROVED BY THE COUNCIL OF INDUSTRIAL DESIGN

evised and improved dial retains the
r and well-known trigger action that has
reliable service in G.E.C. dials for many years.
ain new features of the dial include:

ulded, hard-plastic case
ing the former metal case

stic finger plate available in seven
ive colours to match the cases of the
Gecophone and the B.P.O. Telephone No. 706

struction card cover fits
with the finger plate so
a pleasing appearance

lum contacts are fitted
impulsing springs



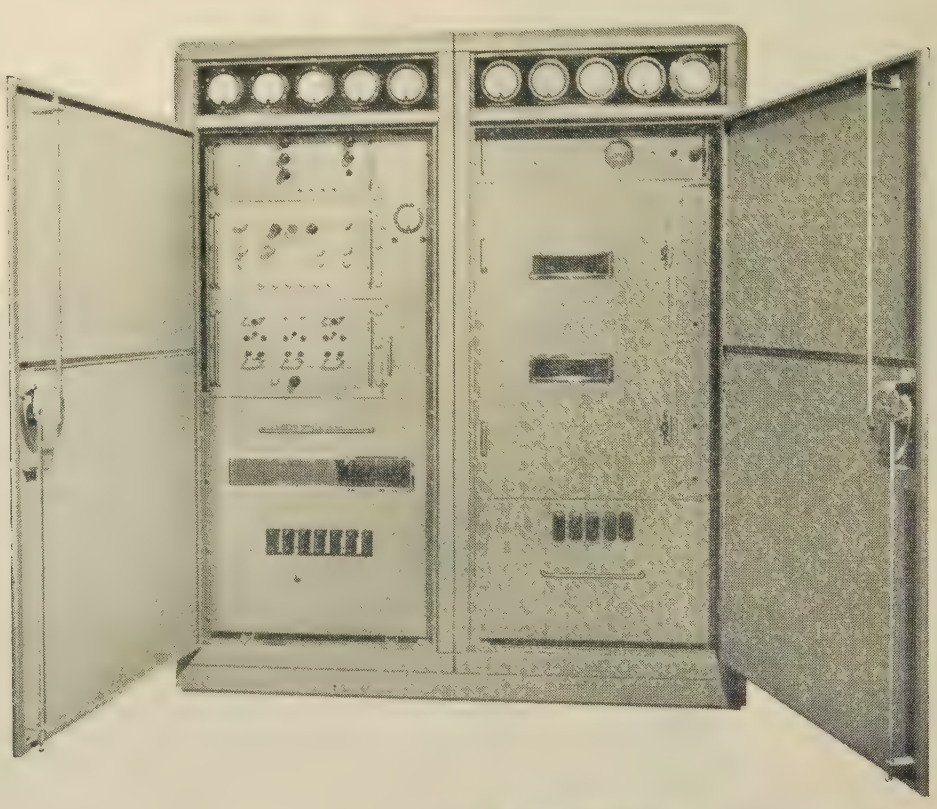
ew dial is fitted in the New Gecophone
e B.P.O. Telephone No.706, which is approved
British Council of Industrial Design.
se applications the dial number ring,
is coloured to match the telephone handset,
blank as a number ring is fitted
case of the telephone. However, the dial
so be fitted to any telephone
ses the previous G.E.C. trigger dial.
ese uses the numbers are featured on the
mber ring in the conventional manner.
required, plastic covers are supplied
ect the mechanism from dust and damage.

dial has been accepted as standard by the British Post Office
has been designated the P.O. Dial No. 21

G.E.C.

THE GENERAL ELECTRIC COMPANY LIMITED OF ENGLAND
Telephone, Radio and Television Works, Coventry, England.

GEC48



THIS PACKAGE CONTAINS:

5 kilowatts R F power immediately available from your 500 watt transmitter

RACAL Engineering Limited

- S.S.B., C.W., D.S.B. or F.S.K.—2.5 to 25 Mc/s
- Easily coupled to existing 300–500 watt transmitter
- Fully-automatic tuning, rapid channel-changing
- Remote control facilities

The Racal TA.84 Linear Amplifier

is ideal for long-distance ground/air links, marine shore stations, meteorological stations, fixed and semi-mobile military uses, national security networks and medium-distance R/T circuits—indeed for nearly all radio links. It gives a power amplification of at least 10 times with an R.F. output of 5 kW P.E.P. from an R.F. input of 500 W or 3 kW R.M.S. Its simplicity of operation and tuning accuracy demand little or no skilled attention in the transmitting station. It is designed to comply fully with Joint-Services Specification K.114 (Protected Ground Equipment), soundly constructed and fit for the most adverse climatic and operating conditions. Full specifications and details on request.

RACAL

RACAL ENGINEERING LIMITED

WESTERN ROAD, BRACKNELL, BERKS, ENGLAND Tel.: Bracknell 941 'Grams/Cables: RACAL BRACKNELL BERKS

Overseas agents in most territories



WHAT HAS THE **7182** **S BAND MAGNETRON** IN COMMON WITH THIS CAT?



A good question! According to legend, a cat has nine lives—something that has yet to be proved. The 7182, on the other hand, has a *proved* life 8 to 10 times greater than any similar S-Band Magnetron.

The 7182 retains its remarkable stability and reliability during a life of 10,000 hours, and is one of a series of Magnetrons now in production providing peak powers of 5MW. These Magnetrons operate at voltages and current densities usually associated with Magnetrons rated at a fraction of the power output.

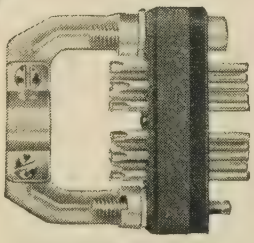
A parallel development in L-Band ensures that attainments in this field equal the phenomenal successes already achieved with S-Band Magnetrons.

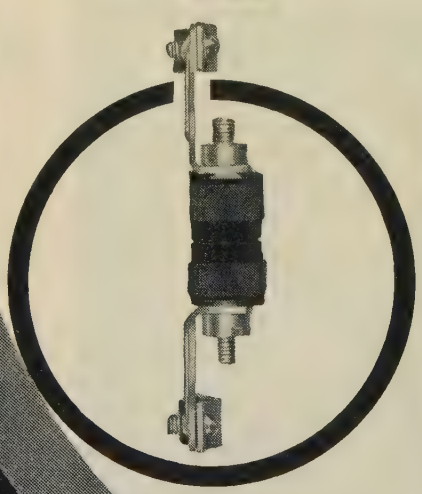
'ENGLISH ELECTRIC'
Agents throughout the world

ENGLISH ELECTRIC VALVE CO. LTD. *CHELMSFORD, ENGLAND*
Telephone : Chelmsford 3491

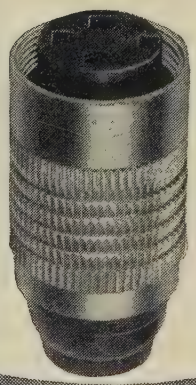
AP134

a firm GRIP on complex connection









From their wide selection of plugs and sockets, MODAC offer a standard range of rectangular and circular miniature connectors, to which is soon to be added a sub-miniature range with combinations of 10, 16, 28 and 34 contacts.

The main feature of this advanced design is the independently sprung, multi-point contact which provides self-cleaning, low contact resistance and is ideal under conditions of vibration.

MODAC is equipped for the design of connectors to customers' specifications and to supply any of their range as cable assemblies.

They are more than willing—at any time—to come to grips with your general wiring problems.

Modac miniature and sub-miniature connectors

'Modac' is a Registered Trade Mark

for further information and a copy of the 'Modac' catalogue, please contact:

Wiring & Connectors Division
The Plessey Company Limited

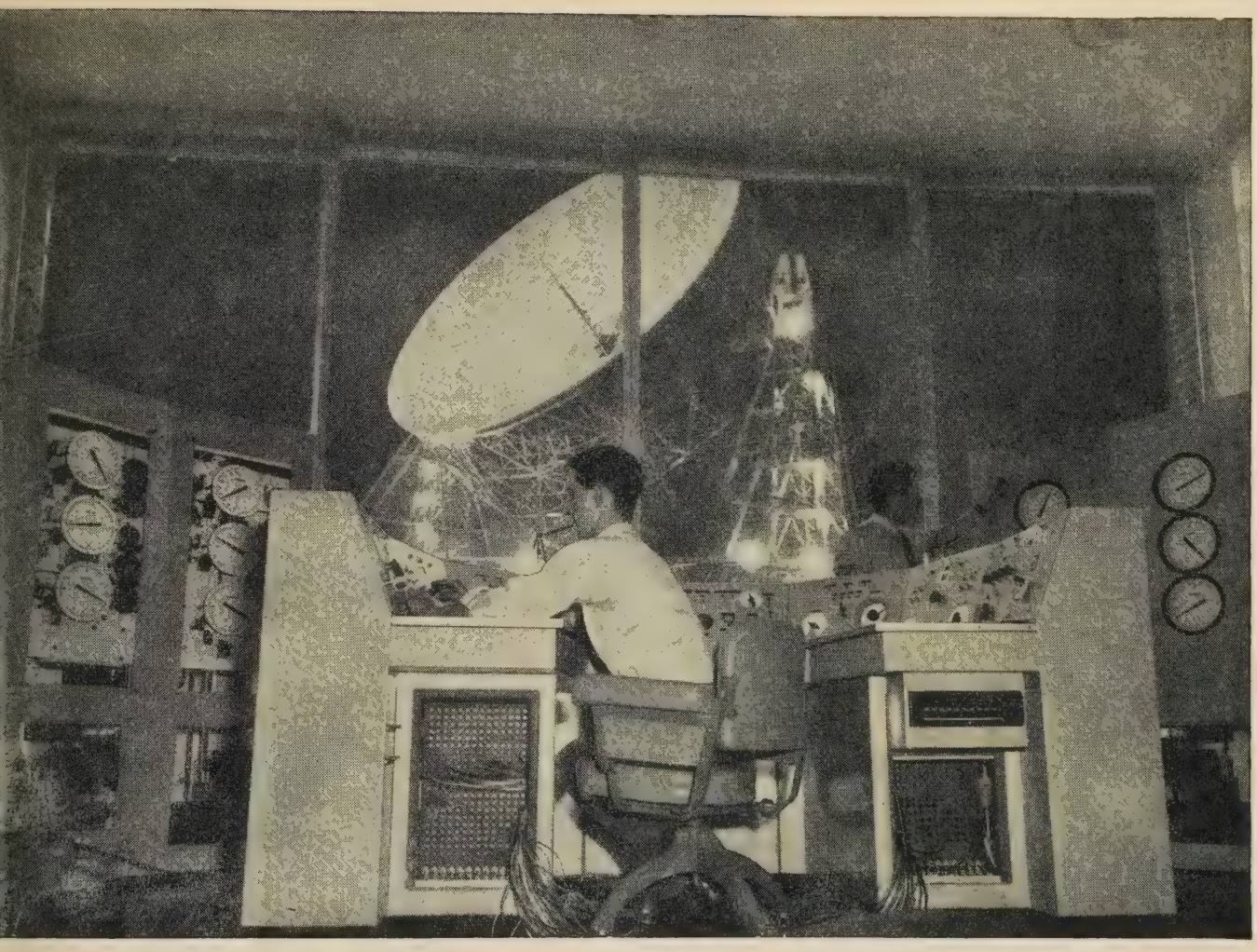
CHENEY MANOR · SWINDON · WILTS · Telephone: SWINDON 6251

Agents for British Commonwealth Countries: Plessey International Limited · Ilford · Essex · Tel: Ilford 3040





MAL 2

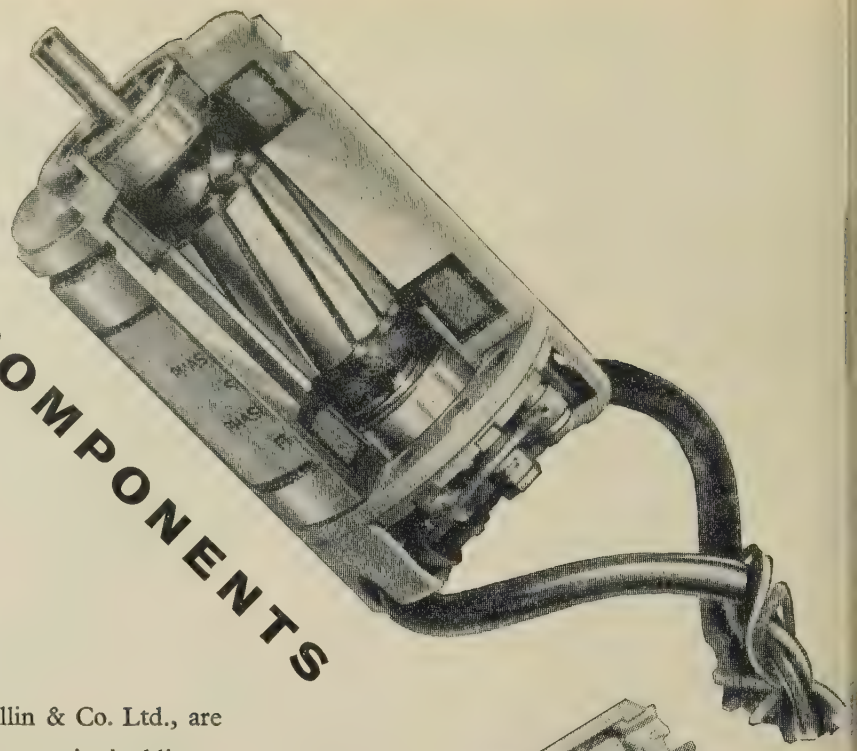


Symbolic of this age when science fiction is scarcely written before it becomes fact, the radio telescope at Jodrell Bank emerges as man's first major link with a limitless universe. For this vast undertaking of mechanical and electronic precision only materials and components of the highest quality were used, and we are understandably proud to record that Dubilier capacitors and resistors were chosen by Messrs. Dunford & Elliott (Sheffield) Ltd., as suppliers of the control equipment.

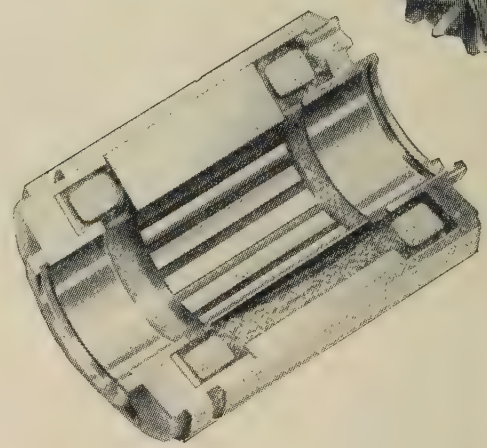
DUBILIER

DUBILIER CONDENSER CO. LTD., DUCON WORKS, VICTORIA RD., NORTH ACTON, LONDON, W.3. Tel: ACOrn 2241. Grams: Hivoltcon Wesphone London
DN 226

Araldite IN SERVOCOMPONENTS



The Synchro units shown, made by R. B. Pullin & Co. Ltd., are sectioned to show how the stators are integrally cast in Araldite to provide maximum protection against the effects of extremes of temperature, humidity and vibration. The excellent machining properties of Araldite make possible a straight-through bore technique, which eliminates errors in alignment and also permits the use of the smallest possible air gap between rotor and stator. High insulation and dielectric strength, remarkable adhesion to metals, and negligible shrinkage on curing make Araldite eminently suitable for use in the construction of precision electrical equipment.



Araldite epoxy resins are used—

- **for casting high grade solid electrical insulation**
- **for impregnating, potting or sealing electrical windings and components**
- **for producing glass fibre laminates**
- **for producing patterns, models, jigs and tools**
- **as fillers for sheet metal work**
- **as protective coatings for metal, wood and ceramic surfaces**
- **for bonding metals, ceramics, etc.**



This photograph shows an A.E.W. electric oven, capable of maintaining temperatures within close limits, as used by R. B. Pullin & Co. Ltd. for curing the Araldite-filled stators.

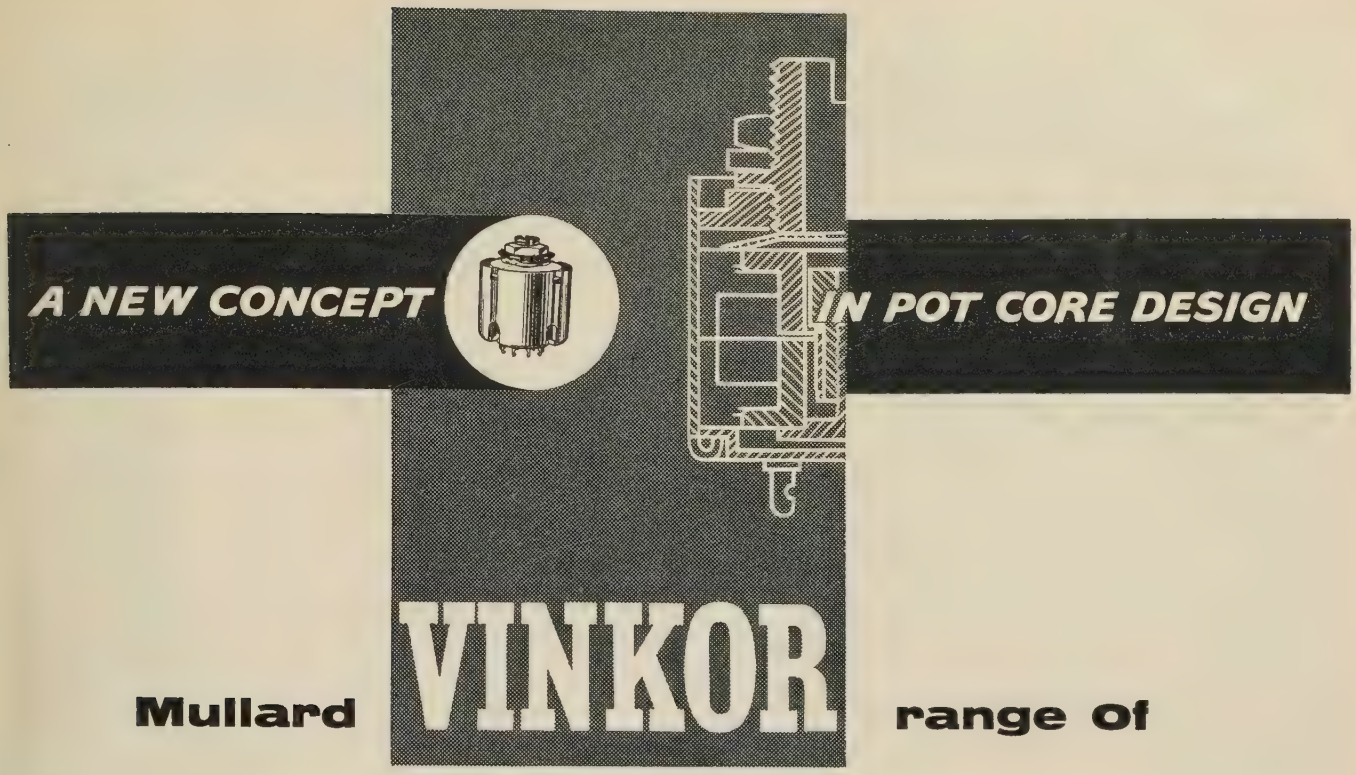
Araldite *epoxy resins*

Araldite is a registered trade name

CIBA (A.R.L.) LIMITED

Duxford, Cambridge Telephone: Sawston 2121

AP 463



Mullard range of adjustable pot cores gives you *outstanding* advantages

- Wide range of sizes
- Easily assembled
- Close tolerance permeability
- Precise and easy inductance adjustment
- Stability
- Single hole chassis mounting

Mullard Vinkors are the most efficient adjustable pot core assemblies commercially available. In addition to high performance, they have the distinct advantage of close tolerance permeability, thus enabling designers to precalculate to within ±3% the inductance of the core when wound. Final adjustment, taking into account normal capacitor tolerance, can be easily effected to an accuracy of better than 0.02%, by means of a simple self-locking device built into the core.

Write today for full details of the wide range of Vinkors currently available.

Mullard

VINKOR POT CORES

MULLARD LIMITED · COMPONENT DIVISION · MULLARD HOUSE · TORRINGTON PLACE · LONDON W.C.1

MC280

you are invited to exhibit . . .

. . . at the



International

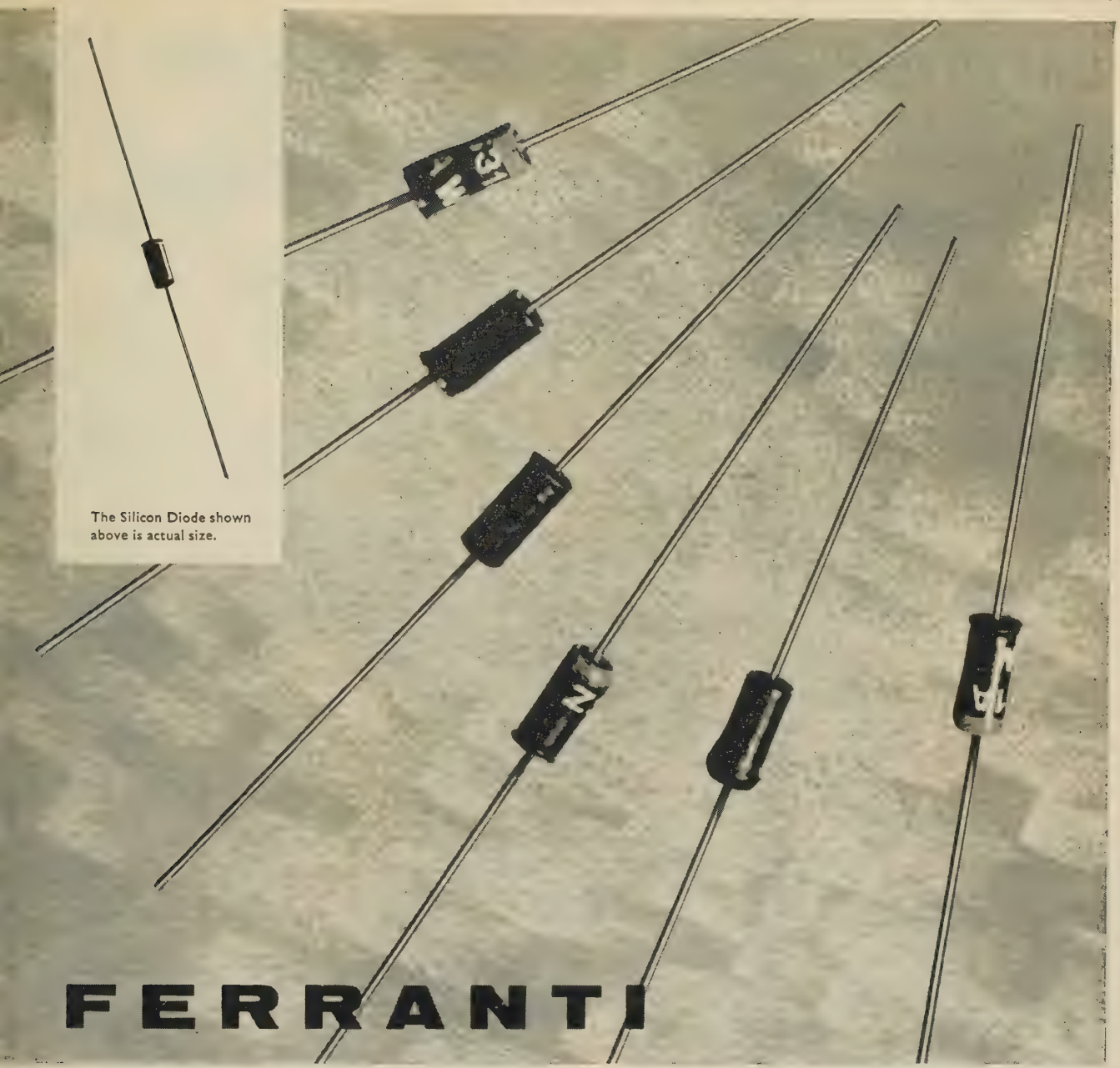
MEDICAL ELECTRONICS EXHIBITION

OLYMPIA LONDON - 21-27 JULY, 1960

In conjunction with the Third International Conference on Medical Electronics organized by the Electronics and Communications Section of The Institution of Electrical Engineers, in association with The International Federation for Medical Electronics.

Full details from:-

Industrial Exhibitions Ltd., 9 Argyll Street, London, W.1. Telephone: Gerrard 1622



The Silicon Diode shown above is actual size.

FERRANTI

Silicon Diodes ZS30 series

for MINIATURIZED Circuitry

SPECIAL FEATURES

- AUTOMATIC WIRING TECHNIQUES
- HIGH POWER TO SIZE RATIO
- HIGH TEMPERATURE OPERATION
- HIGH FORWARD CONDUCTANCE
- HIGH RECTIFICATION EFFICIENCY
- RUGGEDISED CONSTRUCTION

Ferranti Miniaturized Circuit Diodes are designed for automatic wiring techniques. These diodes, in addition to being run at maximum ratings for forty-eight hours, are rigorously tested to satisfy the following conditions:

Vibration	Grade I (Aero Engines)
Shock	> 500 g.
Humidity	Class H1
Temperature Range	−70°C to + 135°C.

P.I.V. Range 50–400 Volts: Max. Mean Rectified Current 500 mA

FERRANTI

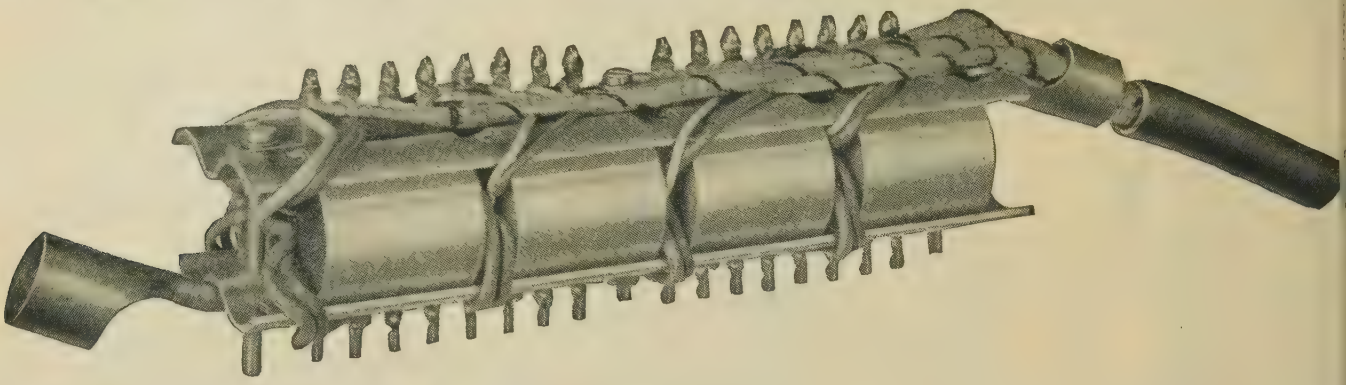
FERRANTI LTD · GEM MILL · CHADDERTON · OLDHAM · LANCS · Tel: MAin 6661

LONDON OFFICE: KERN HOUSE, 36, KINGSWAY, W.C.2. Telephone: TEMple Bar 6666

FERRANTI SILICON SEMICONDUCTOR DEVICES

FE 194

Improved splice loading with the L219

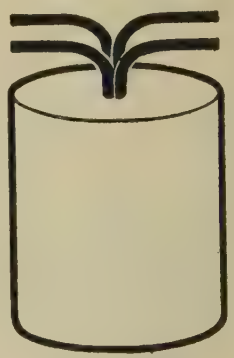


The new economical loading coil

Arising from the increasing demand for a smaller coil which can be employed in splice loading, the L219 has been developed. In the design Mullard Equipment Limited were assisted by their own production experience and information given by overseas users. The result is a simple, low cost component (to grade 3 spec.) suitable for small or large splice loading units.

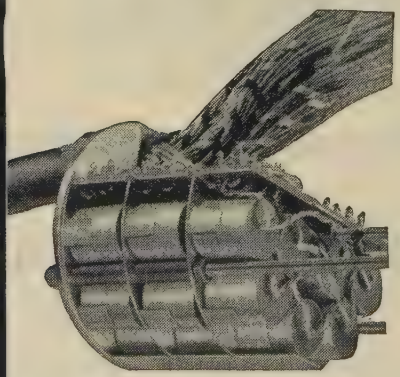
Smaller construction

By using a new grade of Ferroxcube pot core the overall volume of the coil is considerably reduced. The coil is resin sealed in a small cylindrical aluminium canister ensuring complete protection from climatic effects. The windings of the coils are brought out on flying leads.



LIFE SIZE COIL
L219

Permits smaller splices



Key factors in this development are the clamping arrangements which, with the new coil, permit much smaller splice housing. On small cables, coils are mounted lengthways in pairs with great compactness. For larger cables, coils are mounted radially, each mounting plate accommodating up to seven coils. Clamping plates, coils, etc. can be supplied as kits.

Please write for full details of these new loading coils

PRODUCTS OF
MULLARD EQUIPMENT LIMITED
A COMPANY OF THE MULLARD GROUP

MULLARD EQUIPMENT LIMITED

A Company of the Mullard Group

Mullard House · Torrington Place · London W.C.1 · Telephone: Langham 6633

ME614a



35 YEARS OF 'KNOW HOW' BEHIND EVERY TERMINAL

We make millions of terminals in a most comprehensive range of types and sizes. In many cases, collar assemblies are interchangeable, thus making possible a wide variety of combinations. Where our range—extensive as it is—does not provide for your particular requirements, we are always prepared to 'make to measure'.

We are specialists in components. This is no empty phrase: it means that we have been engaged in the design and manufacture of precision-made components for some thirty-five years, and at Enfield we devote a very great deal of our time and space to research, forever seeking new and better methods of production, whilst ensuring that every component maintains our high standard of performance.

Most "Belling-Lee" products are covered by patents or registered designs or applications therefor.

BELLING & LEE LTD
GREAT CAMBRIDGE ROAD, ENFIELD, MIDDX., ENGLAND
Telephone: Enfield 3322 · Telegrams: Radiobel, Enfield

Regd.

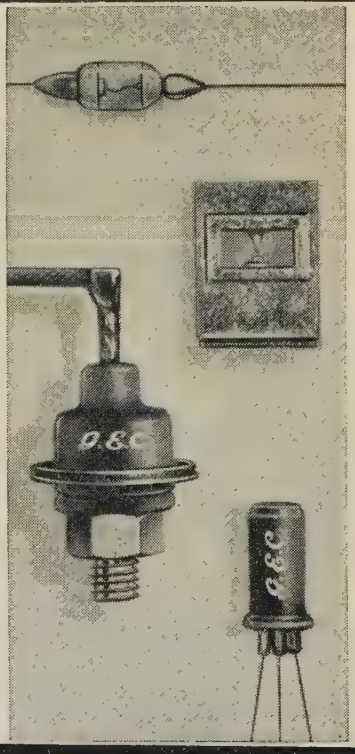
TERMINALS · PLUGS & SOCKETS · THERMAL DEVICES
FUSES · INTERFERENCE FILTERS · RECEIVING AERIALS

*b

3Q PRODUCTION

G.E.C. original processes now achieve QUALITY QUANTITY QUICKLY

3Q production—Quality, Quantity—Quickly! That's the ideal production combination for semiconductors we've now achieved with G.E.C.-originated manufacturing processes. And it's because these new processes have so revolutionised our production that you can be sure of getting the G.E.C. devices you want—when you want them! We offer you the widest range in the country—at really competitive prices too!



G.E.C. has been first with all these achievements!

Microwave detectors and mixers

Point contact diodes

Cold-welded copper-to-copper sealing

Copper sealing glass

85°C. junction temperature for germanium

12 amp. power transistor

Watch for future advertisements describing these and other new G.E.C. developments

QUALITY · QUANTITY · *QUICKLY*

G.E.C. SEMICONDUCTORS

For information on the range of G.E.C. semiconductor devices please write to: G.E.C. Semiconductor Division, School Street, Hazel Grove, Stockport, Cheshire, or in London area, phone: TEMple Bar 8000, Extension 10.

MARCONI VACUUM-TUBE VOLTMETER TF1041B- *for u.h.f precision*



For unprecedented performance and versatility

The Marconi TF 1041B is a versatile, precision voltmeter offering an unprecedented performance in the measurement of a.c. voltage, d.c. voltage, and resistance.

For a.c. measurement the frequency range extends from 20 c/s to 1,500 Mc/s and the voltage range from 25 mV to 300 volts; input capacitance is only 1.5 μμF and input resistance is 0.5 MΩ at 10 Mc/s.

The d.c. measurement range is 10 mV to 1,000 volts and input impedance is 100 MΩ; the meter can be switched to give forward deflection for positive or negative voltages, or to allow centre-zero measurements.

Resistances can be measured from 0.02Ω to 500 MΩ.

Optional accessories include multipliers for up to 2 kV a.c. and 30 kV d.c., a T-connector for measurements on co-axial lines, and a wide-band coaxial dummy load. Send for leaflet K163.

MARCONI INSTRUMENTS

AM & FM SIGNAL GENERATORS · AUDIO & VIDEO OSCILLATORS
FREQUENCY METERS · VOLTMETERS · POWER METERS
DISTORTION METERS · FIELD STRENGTH METERS
TRANSMISSION MONITORS · DEVIATION METERS
OSCILLOSCOPES, SPECTRUM & RESPONSE ANALYSERS
Q METERS & BRIDGES

Please address enquiries to MARCONI INSTRUMENTS LTD. at your nearest office:

London and the South: Marconi House, Strand, London, W.C.2 Telephone: COVent Garden 1234

Midlands: Marconi House, 24 The Parade, Leamington Spa Telephone: 1408

North: 23/25 Station Square, Harrogate Telephone: 67455

Export Department: Marconi Instruments Ltd., St. Albans, Herts. Telephone: St. Albans 56161

TC 163



Marconi in Telecommunications



The post and telegraph authorities of more than 80 countries use Marconi equipment

MARCONI

COMPLETE COMMUNICATION SYSTEMS

SURVEYED · PLANNED · INSTALLED · MAINTAINED

MARCONI'S WIRELESS TELEGRAPH COMPANY LIMITED, CHELMSFORD, ESSEX, ENGLAND

M.1

The most advanced voltage reference tube in the world . . .

the NEW Mullard 83A1

or technical details of the 83A1 fill in and post this coupon

Mullard

GOVERNMENT AND INDUSTRIAL VALVE DIVISION

Send me technical details of the Voltage Reference Tube 83A1

Name

Position

Company

Address

..... P.I.E.E.

Mullard

Mullard Limited, Mullard House, Torrington Place, London, W.C.1

MVT374

DELAYED PULSE AND SWEEP GENERATOR



A versatile pulse generator designed to meet the need for a comprehensive instrument covering a wide range of pulse work. Four main facilities are provided: a pre-pulse, a main pulse delayed on the pre-pulse, a negative going sawtooth and a fast rising pulse formed from a pure line.

BRIEF SPECIFICATION

Period

Continuously variable from 0·9μsec to 1·05sec i.e. 0·95c/s to 1·1Mc/s. Accuracy ±5%.

Pre-pulse

40mμsec. 8V peak in 75Ω, positive going.

Main pulse

Width: Variable from 0·09μsec to 105msec ±5%.

Amplitude: Control gives 4:1 attenuation of each of four maximum outputs as follows:

5V max in	75Ω	rise time 10mμsec
10V max in	150Ω	rise time <20mμsec
25V max in	600Ω	rise time <40mμsec
50V max in	1000Ω	rise time 50mμsec

Polarity: Positive or negative going.

Accuracy: ±2%.

Delay

Conclusion of pre-pulse to advent of main pulse, delay variable from 0·09μsec to 105msec. Accuracy ±5%.

Sweep

D.C. coupled negative going sawtooth same width and delay as main pulse.
15V peak max.

Cable pulse

Obtained from short circuited pure line.
One positive and one negative going pulse coincident with main pulse.
25mμsec wide 3V max in 75Ω, rise time <8mμsec.

Sync, trigger or single shot facilities provided.
Full data available on request.

CINTEL

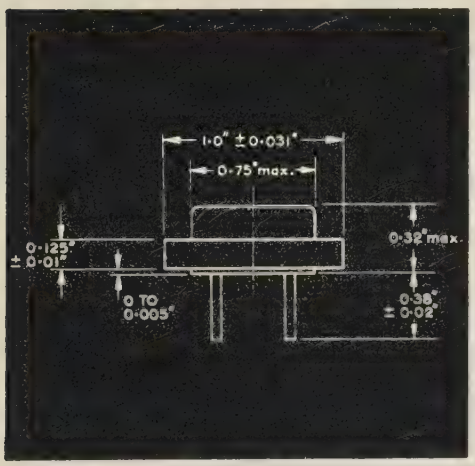
RANK CINTEL LIMITED
WORSLEY BRIDGE ROAD · LONDON · SE26
HITHER GREEN 4600

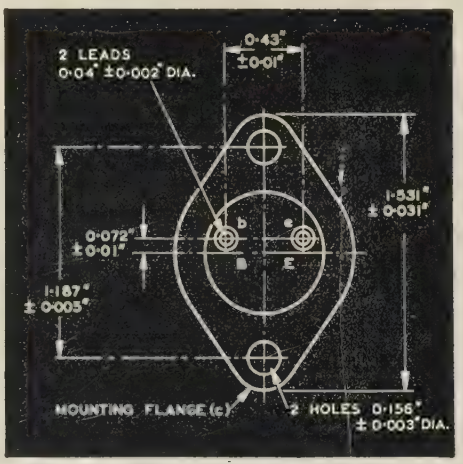
Sales and Servicing Agents: Atkins, Robertson & Whiteford Ltd., Industrial Estate, Thornliebank, Glasgow.
McKellen Automation Ltd., 122 Seymour Grove, Old Trafford Manchester, 16. Hawnt & Co., Ltd., 59 Moor St., Birmingham, 4.

Audio power output transistors

TYPES XC141 and XC142

These germanium p-n-p alloy junction transistors are designed for use in Class A and Class B power output stages of audio frequency amplifiers. Full particulars of these and other Ediswan Mazda semiconductor devices will be sent gladly on request. If you wish to be kept up to date with the latest developments in this field, please ask us to add your name to our semiconductor mailing list.





MAXIMUM RATINGS (Absolute Values)	XC141	XC142
Peak collector to base voltage (volts)	– 40	– 60
Peak collector to emitter voltage, emitter non-conducting (volts)	– 40	– 60
Peak collector to emitter voltage, emitter conducting (volts)	– 32	– 32
D.C. Emitter to base voltage (volts)	– 12	– 12
Peak collector current (amps)	– 3·0	– 3·0
D.C. Collector current (amps)	– 1·5	– 1·5
Collector dissipation (mounting flange temperature 80°C) (watts)	11	11

EDISWAN SEMICONDUCTORS
MAZDA

Associated Electrical Industries Ltd
Radio and Electronic Components Division
PD 15, 155 Charing Cross Road, London, W.C.2
Tel: GERrard 8660 Telegrams: Sieswan Westcent London

CRCI 5/58

A.T.E constructs for Canada

The A.T.E. Group recently completed their part of an important telecommunication project in Newfoundland and Nova Scotia. The contract was placed by the Canadian National Telegraphs and included ATE high grade channelling equipment for use over microwave radio relay links supplied by Standard Telephones & Cables Ltd.

The newly completed communications route stretches from St. Johns, Newfoundland, to Sydney, Nova Scotia, and the ATE equipment comprises units for an initial 2 supergroups, one of which is fully equipped for 60 circuits. Equipment for one bothway and one unidirectional 10 kc/s broadcast channel has also been supplied.

A.T.E. have also been commissioned to provide equipment for an additional link between Cornerbrook and Deer Lake. This spur will cater for three 12-channel groups working into American 'L' type carrier equipment at Deer Lake.

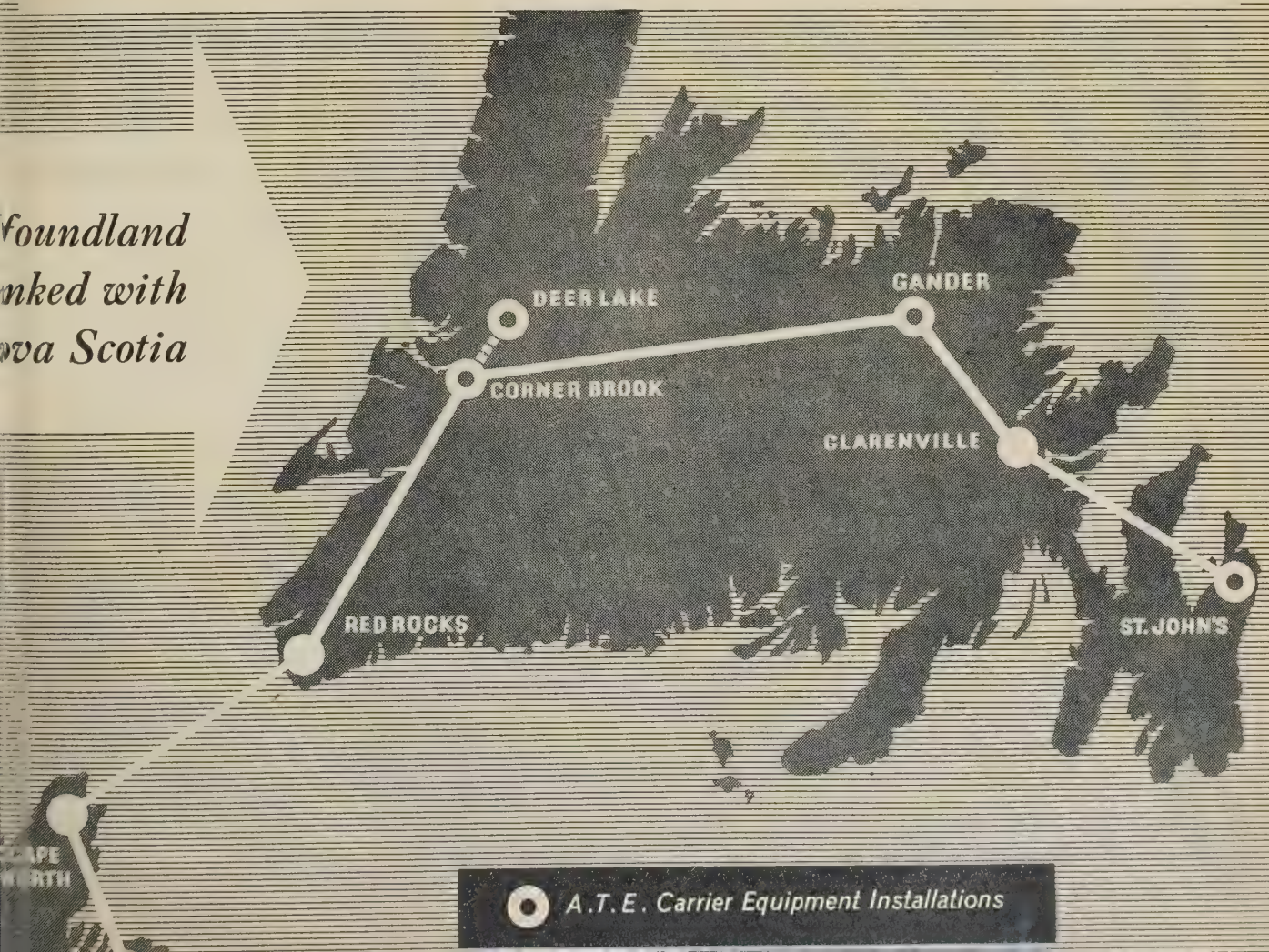
Canada is one of 20 countries in which the ATE Group has recently completed large telecommunication projects.

AUTOMATIC TELEPHONE & ELECTRIC CO. LTD.

STROWGER HOUSE, ARUNDEL STREET, LONDON. W.C.2. TEL: TEMPLE BAR 9262. TELEGRAMS: STROWGER, ESTRAND, LONDON

AT8901

foundland
nked with
va Scotia



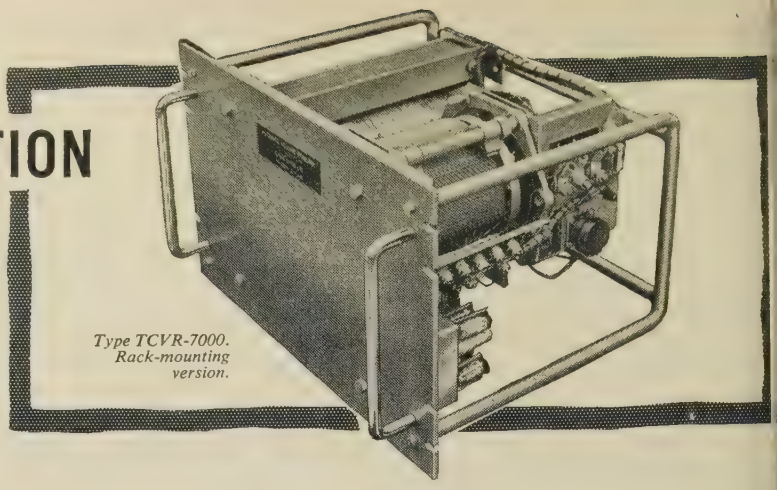
40 VOLTS/SEC AUTOMATIC CORRECTION

—with the type TCVR voltage regulator

The TCVR is a servomechanical automatic voltage regulator having the very high speed of correction of FORTY VOLTS PER SECOND. It provides an *undistorted* output, maintained constant within very close limits (normally ±0.5%) from no-load to full-load, for wide variations in frequency and power factor.

A wide range of models from 1.6 to 12 kVA single-phase, and 4.8 to 36 kVA three-phase, is available, to standard or tropical specification, in cabinets or for rack-mounting. Models are also available in which the output voltage is continuously adjustable over a wide range by means of a panel control. Regulators can be supplied to Services' specification, and special models can be designed to order.

For high-speed, accurate stabilisation *without distortion*—**specify TCVR.**



Type TCVR-7000. Rack-mounting version.

Claude Lyons Ltd.

Other products of Claude Lyons Ltd. Stabiliser Division

BMVR: Motor-driven laboratory and industrial regulators ranging from 1·6 to 29 kVA single-phase, and 4·8 to 87 kVA three-phase. Constancy of output normally ± 0·5% from no-load to full-load. No distortion. Speed of correction 1 Volt/Sec. A great variety of models, standard, tropical and militarised, for all applications.

BAVR: Electronic stabilisers of very high accuracy, and very rapid response, with no moving parts. Input range : – 10% to + 5%, output constancy ± 0.15%. Three sizes : 200, 500 & 1000 VA. Exceptionally useful for control of chemical processes, heating, lighting, etc.

ASR: Automatic step regulators, small, inexpensive, and with sinusoidal output waveform. Two sizes : 1·15 kVA and 2·3 kVA. Input range – 10% to + 5%; output constancy, ±2½%.

ATC: Automatic Tap-Changing Transformers — a development of ASR. Two sizes : 575 VA and 1150 VA. Input range – 20% to + 10% : output constancy, ± 5%. Provide adequate stabilisation for many types of apparatus, at low cost. Also useful as pre-regulators, e.g. in conjunction with BAVR.

We shall be pleased to send you full details of our entire range.

Stabiliser Division **VALLEY WORKS · HODDESDON · HERTS · TELEPHONE HODdesdon 4541–4**

CL/47/E21

What every Engineer should know.... about MULTICORE solder

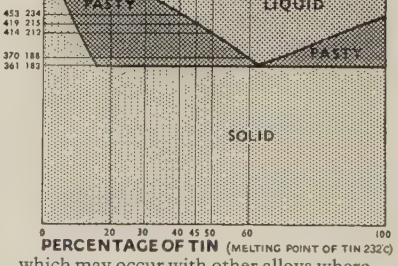


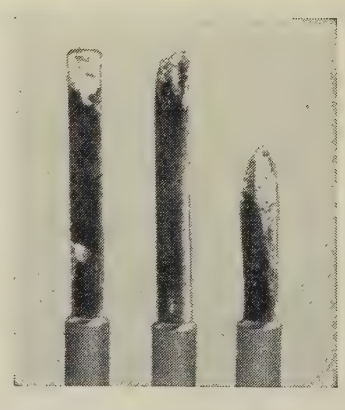
ERSIN MULTICORE 5-CORE SOLDER

The A.I.D. approved type 362 flux, incorporated in Ersin Multicore 5-Core Solder, is very effective on heavily oxidised surfaces and often allows the use of a lower tin content alloy. Ersin Multicore 5-Core Solder is supplied on 7 lb. reels in 9 standard gauges and 6 alloys. Even gauges from 24-34 s.w.g. are available in 2 alloys on 1 lb. and ½ lb. reels.

CONSTITUTION OF ALLOYS OF ERSIN MULTICORE SOLDER

The diagram shows that all the standard alloys of Ersin Multicore Solder have a plastic range, i.e., on heating they are pasty between the solid and liquid states. Practical experience has shown that there are advantages in having a plastic range, e.g. for tag jointing where the use of 60/40 alloy obviates fractures which may occur with other alloys where there is slight vibration while the solder is setting solid.





PRINTED CIRCUITS

Leaflet P.C.L. 101 gives full details of a complete soldering process developed by Multicore Laboratories for printed circuits.



SAVBIT TYPE 1 ALLOY *Made under sole British Licence of Patent No. 721,881.*

Savbit Type 1 alloy was developed after extensive research in the Multicore Laboratories into the main causes of bit wear. It incorporates a small percentage of copper which prevents absorption of copper from the bit into the solder alloy. After prolonged tests, it was found that the life of solder bits was increased by up to 10 times. The speed of soldering is not affected.

SPECIAL ALLOYS

4 special alloys can be supplied for particular purposes:

Comsol with a high melting point of 296°C.

T.L.C. alloy with a melting point of 145°C is made from tin, lead and cadmium.

P.T. (Pure Tin) alloy, with a melting point of 232°C, is lead-free.

L.M.P. alloy, with 2% silver content, which melts at 179°C for silver-coated components.

PUBLICATIONS

Laboratory engineers and technicians are invited to write on their company's letter-heading for the latest edition of Modern Solders. It contains data on melting points, gauges, alloys, etc.

MULTICORE SOLDERS LTD., MULTICORE WORKS, HEMEL HEMPSTEAD. HERTS.

Tel: Boxmoor 3636 (4 lines). Grams and Cables: Multicore Hempstead.

Semiconductors

COMPUTER TRANSISTORS



The Semiconductors range of Computer Transistors, designed and tested to the special requirements of computer engineers, is the key to a new order of computer speed and reliability. Overall reliability is further increased by making possible a substantial reduction in the number of associated components.

The two types of Silicon Alloy Transistor shortly going into production will make it possible to extend this high-speed computer performance into **ambient** temperatures well above 100°C. **Samples are available now.**

	TYPE	DESCRIPTION	RISE TIME millimicroseconds	Vc max	Ic max
HIGH-SPEED LOW-LEVEL SWITCHING GERMANIUM	SB 344 SB 345	General purpose transistors for conventional logic circuits.	50	5v	5mA
	SB 240	Designed for directly coupled circuits. Controlled input, saturation and hole storage characteristics.	30	6v	15mA
	MA 393	High gain transistor for high-speed driving of parallel circuits.	30	6v	50mA
	2N 501	Ultra-high speed transistor with controlled input and saturation characteristics.	10	12v	50mA
HIGH-SPEED LOW-LEVEL SWITCHING SILICON	SA 495	General purpose 10Mc/s transistor for conventional logic circuits.	100	25v	50mA
	SA 496	15Mc/s transistor for directly coupled circuits. Saturation resistance typically 10 ohms. Controlled input and hole storage characteristics.	80	10v	50mA
CORE DRIVING GERMANIUM	2 N 597 2 N 598 2 N 599	min $f\alpha$ 3Mc/s min $f\alpha$ 5Mc/s min $f\alpha$ 12Mc/s 250 mW high frequency alloy transistors with high gain and low saturation resistance	400 * 250 * 100 *	20v 20v 20v	400mA 400mA 400mA
	2 N 600 2 N 601	min $f\alpha$ 5Mc/s min $f\alpha$ 12Mc/s 750 mW versions of 2 N 598 and 2 N 599. Peak current 3 amps.	250 * 100 *	20v 20v	400mA 400mA

* *rise time to 400mA*

Full technical details and applications assistance available on request.

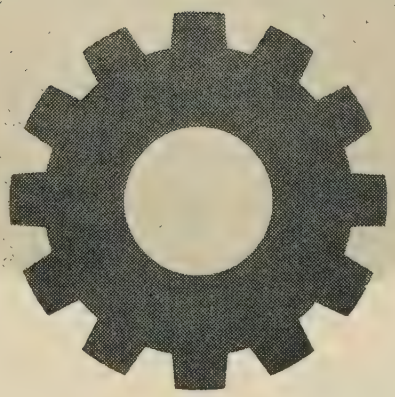
Semiconductors Limited

CHENEY MANOR
SWINDON · WILTS
TELEPHONE: SWINDON 6421/7

S.C.10

PHILIPS

In Science and Industry alike . . .



among technicians, manufacturers and those engaged in the sale of electrical products — as well as among the public at large, the Philips emblem is accepted throughout the World as a symbol of quality and dependability.

PHILIPS ELECTRICAL LTD

Century House · Shaftesbury Avenue · London · WC2

Radio & Television Receivers · Radiograms & Record Players · Stereo & Hi-Fi Equipment · Tape Recorders · Dictation Machines · Gramophone Records · Tungsten, Fluorescent, Blended & Discharge Lamps & Lighting Equipment · "Photoflux" Flash Bulbs · "Philishave" Electric Shavers · Health Lamps · Hearing Aids · Electric Blankets · Domestic Electrical Appliances X-Ray Equipment · High Frequency Heating Generators · Electro-Medical Apparatus · Arc Welding Plant & Electrodes · Electronic Measuring Instruments · Magnetic Filters · Sound Amplifying Installations · Cinema Projectors

THE JOURNAL OF

The British Nuclear Energy Conference

The Institution of Civil Engineers
The Institution of Mechanical Engineers
The Institution of Electrical Engineers
The Institute of Physics
The Institution of Chemical Engineers
The Institute of Metals
The Iron and Steel Institute
The Institute of Fuel
The Joint Panel on Nuclear Marine Propulsion

PUBLISHED JANUARY, APRIL, JULY, OCTOBER

The Journal contains papers and discussions on the applications of nuclear energy and ancillary subjects

ANNUAL SUBSCRIPTIONS:

MEMBERS 30/- post free
NON-MEMBERS 60/- post free

Full particulars are available from

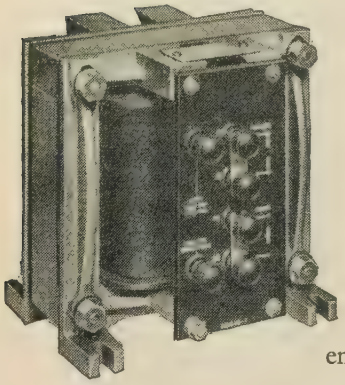
The Secretary • B.N.E.C. • 1-7 Great George Street • London • SW1



MARCONI'S WIRELESS TELEGRAPH COMPANY LIMITED, CHELMSFORD, ESSEX, ENGLAND
M.8

ZENITH
(REGD. TRADE-MARK)

Transformers



Open shell type transformers from 5 VA up to 5 kVA to B.S.S. 171/36, core type single phase and three phase, air cooled and oil immersed, are included in our range. May our engineers assist your designer to solve that "one off" problem?

The ZENITH ELECTRIC CO. Ltd.

ZENITH WORKS, VILLIERS ROAD, WILLESDEN GREEN
LONDON, N.W.2

Telephone: WILlesden 6581–5 *Telegrams:* Voltaohm, Norphone, London

MANUFACTURERS OF ELECTRICAL EQUIPMENT
INCLUDING RADIO AND TELEVISION COMPONENTS

ADCOLA (Regd. Trade Mark)

Soldering Instruments

ILLUSTRATED

PROTECTIVE SHIELD (CAT. No. 68)

$\frac{1}{8}$ in. BIT MODEL (CAT. No. 70)

Primarily developed for the TRANSISTOR & ELECTRONIC ERA.

Possessing the sharp heat essential for the quick jointing of Transistors, Resistors, etc., thereby avoiding damage to the equipment from heat transfe rence



Cover all requirements for thorough solder jointing in all the fields of **TELECOMMUNICATIONS**

Fully Insulated Elements

Suited to daily use for bench line production

MANUFACTURED IN ALL VOLT RANGES

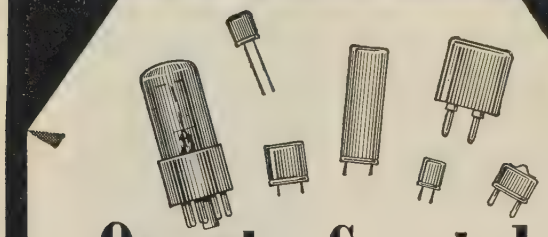
British and Foreign Pats. Reg. designs, etc.

For further information apply Head Office:

ADCOLA PRODUCTS LTD.
GAUDEN ROAD
CLAPHAM
HIGH STREET
LONDON, S.W.4

Tel.: MACaulay 4277 & 3101

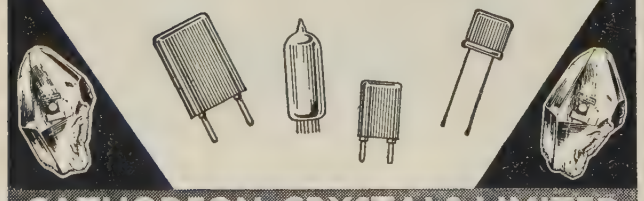
CATHODEON



Quartz Crystals

400c/s to 100M/s

Quick Deliveries
Competitive Prices
E.I.D. and A.R.B. APPROVED



CATHODEON CRYSTALS LIMITED

LINTON CAMBRIDGE ENGLAND TEL: LINTON 501

TRANSISTORISED AUTOMATIC VOLTAGE REGULATORS

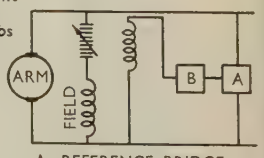




Model shown is for the control of a 28 Volt D.C. generator for use on aircraft.

PATENTS PENDING

Regulation closer than ± 1% between extremes of temperature from –60°C to +70°C
Speed of response 50/60 milliseconds.
For industrial purposes at normal ambient temperatures regulation within ± 0·5%.
Dimensions 5" × 6" × 5½" high. Weight 4lbs

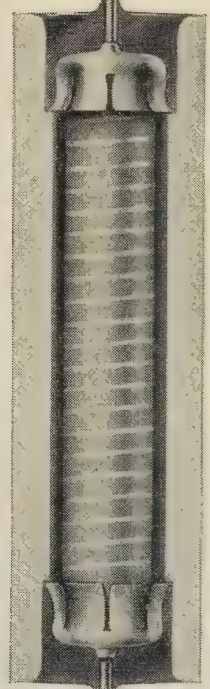


A=REFERENCE BRIDGE
B=TRANSISTOR AMPLIFIER

NEWTON DERBY

NEWTON BROS. (DERBY) LTD.
ALFRETON ROAD · DERBY
PHONE: DERBY 47676 (4 LINES) GRAMS: DYNAMO, DERBY
London Office: IMPERIAL BUILDINGS, 56 KINGSWAY, W.C.2

Fully Protected HIGH STABILITY Resistors



The Inside Story

There is, of course, a lot of "know-how" in the manufacture of a high stability resistor element to the exacting characteristics demanded of a high stability resistor, but, having made the element, two problems remain. How to protect the smoke thin resistance film from damage in transit, in handling, and in assembly, and how to isolate the element from contact with paints, lacquers, and other finishes, all of which have a tendency to pull the film under extremes of temperature and humidity.

In the Erie high stability resistor these two problems have been solved very simply and with complete effectiveness by the encapsulation of the element in the ceramic insulating tube, cement sealed at the ends, proven on billions of Erie solid carbon resistors and ceramic dielectric capacitors in use throughout the world. This tube obviously affords complete protection from all manner of physical damage, and, as can be seen from the illustration, the counter bore at either end and the close affinity between the counter bore and the caps of the element ensures that the element is supported by the caps clear of the inner bore of the tube, and there is thus no contact whatsoever with any material that might prove harmful.

Only Erie high stability resistors are protected in this way, and that is why they are found in all equipment where robustness and reliability under all conditions must be allied to first-class performance.

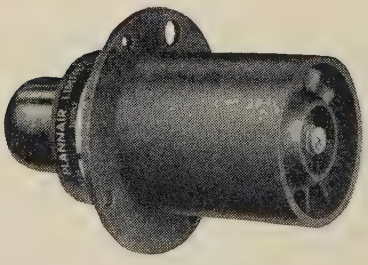
ERIE★ Resistor Ltd

★ Registered Trade Marks

I, HEDDON STREET, LONDON, W.I., Tel. REGent 6432

Factories: Great Yarmouth and Tunbridge Wells, England; Trenton, Canada; Erie, Pa., Holly Springs, Miss. and Hawthorne, Cal., U.S.A.

to keep air moving with the utmost efficiency



PLANNAIR

For thoroughly effective temperature control by the movement of air, design engineers insist on Plannair Blowers—particularly for electronic or aircraft installations where space is at a premium.

Specialist skills and experience explain why Plannair make such a success of making Blowers—consult them from the start and build both their know-how and their Blower right into your equipment.

Certainly if you have—or anticipate—a heating, cooling or ventilation problem, drop Plannair a line or simply ring them; they'll be pleased to give you advice—and specific details about the finest Blowers in the world.

PLANNAIR LIMITED, WINDFIELD HOUSE, EPSOM ROAD, LEATHERHEAD, SURREY · TELEPHONE: LEATHERHEAD 4091

PLA22

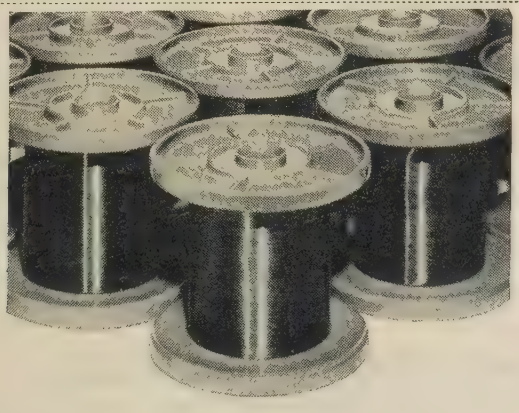
INDEX OF ADVERTISERS

Firm	page
dcola Products Ltd.	xxxii
irmec Ltd.	
r W. G. Armstrong Whitworth Aircraft Ltd.	
ssociated Electrical Industries Ltd.	iii + xxv
utomatic Telephone & Electric Co. Ltd.	xxvi + xxvii
elling and Lee Ltd.	xix
ookhirst Igranic Ltd.	
able Makers Association	
athodeon Crystals Ltd.	xxxii
iba (A.R.L.) Ltd.	xiv
ewhurst and Partner Ltd.	
onovan Electrical Co. Ltd.	
ubilier Condenser Co. Ltd.	xiii
ectro Dynamic Construction Co. Ltd.	vi
LGA Products Ltd.	i
nglish Electric Co. Ltd.	
nglish Electric Valve Co. Ltd.	xi
ie Resistor Co. Ltd.	xxxiii
R.G. Industrial Corp. Ltd.	
ricssons Telephones Ltd.	
erranti Ltd.	xvii
X. Fox Ltd.	
eneral Electric Company Ltd. (M.O. Valves)	ii
eneral Electric Company Ltd. (Semiconductors)	xx
eneral Electric Company Ltd. (Telecommunications)	viii + ix
ternational Medical Electronics Exhbn.	xvi
eorge Kent Ltd.	
odge Plugs Ltd.	
ondon Electric Wire Co. and Smiths Ltd.	xxxv
Claude Lyons Ltd.	xxviii
Marconi Instruments Ltd.	xxi
Marconi Wireless Telegraph Ltd.	xxii + xxxi
Metropolitan Plastics Ltd.	
M-O Valve Co. Ltd.	ii
Mullard Ltd. (Equipment)	xviii
Mullard Ltd. (Components)	xv
Mullard Ltd. (Valves)	xxiii
Multicore Solders Ltd.	xxviii
Newmarket Transistors Ltd.	
Newton Bros. (Derby) Ltd.	xxxii
Oliver Pell Control Ltd.	
Philips Electrical Ltd.	xxx
Plannair Ltd.	xxxiv
Plessey Co. Ltd. (Swindon)	vii + xii
Racal Engineering Ltd.	x
Rank Cintel Ltd.	xxiv
Richard Thomas & Baldwins Ltd.	
Salford Electrical Instruments Ltd.	
Savage Transformers Ltd.	
Semiconductors Ltd.	xxix
Servomex Controls Ltd.	
Siemens Edison Swan (Cables)	
Siemens Edison Swan (Public Telephones)	
Siemens Edison Swan Ltd. (with Redifon)	
Siemens Edison Swan Ltd. (Telecommunications)	
Standard Telephones and Cables Ltd.	
Texas Instruments Ltd.	xxxvi
Telephone Manufacturing Co. Ltd.	iv + v
Zenith Electric Co. Ltd.	xxxii

LEWCOS

Supplied with standard coverings of cotton, silk, rayon, enamel, LEWMEX (synthetic enamel) glass and abestos.

Insulated resistance wires



for resistors in instruments, radio, television, electronic and control apparatus and many other applications.

★ *Available over a range of sizes in VACROM (nickel-chrome) 80/20 or 15% and EUREKA (cupro-nickel). All conform to BSS 115/1954.*

A technical information and advisory service is offered.

THE LONDON ELECTRIC WIRE COMPANY AND SMITHS LIMITED

LEYTON · LONDON · E10



High Performance with Texas Transistors

Smiths 100,000 ft Altimeter



Small Signal Range—2S002-5, 2S014, 3S002, 3S004

Alpha cut-off frequency up to 150 Mc/s · **Collector voltage up to 45V**

Power gain up to 20dB at 70 Mc/s · **Total dissipation up to 150 mW**

Smiths Aircraft Instruments Limited, and many other leading manufacturers, choose Texas transistors. If you would like further details of these or other Texas semiconductors please telephone or write.

STOP PRESS! GERMANIUM POWER TRANSISTORS Available Now Ex Stock **5A 60V** Phone us for details

TEXAS INSTRUMENTS LIMITED

TELEPHONE: BEDFORD 68051 DALLAS ROAD, BEDFORD CABLES: TEXINLIM, BEDFORD

The Institution is not, as a body, responsible for the opinions expressed by individual authors or speakers.
An example of the preferred form of bibliographical references will be found beneath the list of contents.

THE PROCEEDINGS OF 'HE INSTITUTION OF ELECTRICAL ENGINEERS

EDITED UNDER THE SUPERINTENDENCE OF W. K. BRASHER, C.B.E., M.A., M.I.E.E., SECRETARY

)L. 107. PART B. No. 31. JANUARY 1960

The Institution of Electrical Engineers
Paper No. 3133
Oct. 1959
©

INAUGURAL ADDRESS

By SIR WILLIS JACKSON, F.R.S., President.

'THE MAKING OF PROFESSIONAL ENGINEERS'

(*Address delivered before* THE INSTITUTION, *9th October*, 1959.)

INTRODUCTION

When I first entered the Institution building some 30 years o, I soon found myself looking at the tablets mounted at the rance to the lecture theatre on which are inscribed the names those who have been honoured by election to the Presidency our Institution—names which will be linked in perpetuity with evolution of the manifold applications of electricity for the efit of mankind. I am sure that many others must have been ilarly inspired to hope that they might one day be thought rthy to follow them. More recently I have come to regard se names in a somewhat different light—not less as men of ividual distinction, but more as servants of The Institution; as n selected to perform for a short period a special duty symbolic he service to our profession being given by many others whose mes are not similarly displayed; by the members of Council l by the officers and committee members of the specialized tions and Local Centres. These men are the backbone of e Institution. Year by year they support and sustain the sident, and like my predecessors I shall need all the help y can give. I am deeply grateful for the honour which the mbers of The Institution have done me, and I hope I may ve to justify it.

My predecessors have been representative of all branches of profession—of industry and the public services, of its many cializations in power and communication engineering and of foundations in research and education. In general it has n pretty clear which branch it was they individually repre-ted. With me it is not perhaps so clear, for formally and rmally, officially and unofficially, I have dabbled at most. wever, all are likely to identify me with the preparation of ng people for careers in electrical engineering, and it is ut the problems and responsibilities associated with this paration that I want to speak.

When I began to prepare my address I had in mind, as a liminary to the discussion of these matters, drawing some ailed technical comparisons between the situation which ted in electrical engineering when I joined the undergraduate rse at Manchester University in 1922 and that facing the re-ponding present-day students of this subject. In the event I can include only a brief reference to the fantastic pace and scale of the progress which has taken place during the intervening period. In it there have been discovered and developed to the stage of large-scale production a great diversity of new forms of electrical equipment, of which some prominent examples are the pentode, the cavity magnetron, the klystron, travelling-wave tubes and solid-state devices among which the transistor is outstanding; television and radar; systems of short-wave, microwave, and long-distance submarine telephone cable, communication; electronic digital computers; closed-loop servomechanisms and other means of automatic control; new scientific instruments such as the electron microscope and the mass spectrometer, and a whole range of high-energy particle accelerators. In power engineering the changes have been no less outstanding, as exemplified by the development of the Grid, the production of turbo-generators some 10 times larger in capacity and 4 times less in weight per kilovolt-ampere than 35-40 years ago, and the evolution of the nuclear power reactor.

Much of this progress has been made possible by the development of many new magnetic, insulating, semiconducting and structural materials and of improved forms of previously existing ones, and these have also contributed, where applicable, to the considerable changes which have occurred in the types of long-established equipment—dynamos, motors, transformers, instruments, cables, transmission lines, etc.—round which the electrical part of my undergraduate course revolved.

The origins of many of these developments lie in realms of physics, chemistry, mathematics and metallurgy with which I, as a 1925 graduate, and indeed pre-war electrical engineering graduates in general, had the most superficial contact. For example, the concepts of electricity which had been presented to us as basic, and the electric circuit theory we had been taught to use with considerable facility, proved to be quite inadequate to provide an understanding, let alone an ability to initiate a solution, of some of the problems involved. We found ourselves obliged to replace what we thought were the basic principles governing the behaviour of transmission lines and thermionic valves, for example, by a set of more fundamental ones. There became available to us a whole range of new materials, evidently

'60: The Institution of Electrical Engineers

of great potential importance to the progress of our work, the mechanisms of whose special properties we knew virtually nothing. We observed that the trend of our respective specialities could be profoundly affected by what was happening in other specialities with which we had little acquaintance. We found to our embarrassment that we could not be successful specialists unless we were good non-specialists. Few, if any, special courses were available to help us and we dealt with the rapidly changing situation as best we could.

We had to adapt ourselves, not only to the broadening scientific foundations of electrical development, but also to its changing pattern. More and more it involved the handling of major projects, and became a large-scale exercise in team work. It required the professional engineer, and indeed the scientist also, to penetrate more into the manufacturing and organizational fields than previously had been normal. There is nothing new, of course, in the statement that progress in an engineering career requires of the individual more than technical competence. What was new was the rapid increase in the number of men of whom more than technical competence was demanded by the character and the pace of technical development; the increasing occupancy of senior administrative positions by professional engineers and scientists, and the consequential inquiry by these men into means by which their junior colleagues might be prepared more effectively for these wider responsibilities.

Moreover, these wider responsibilities have extended enormously in scope if not in basic character. Engineering concerns the applications of science to the needs of man and society. It is therefore inseparable from humanism. The engineer is up to the neck in human problems whether he likes it or not. He is producing changes in man's mode of life over an extending range and with increasing speed and directness, and raising social problems the consideration of which must form an integral part of his technical and administrative thinking and decisions. It may rightly be said that this also has always characterized engineering activity. What is new in my time, and quite recently, is the explicit recognition and open discussion of the humanistic aspect of technology, and the deliberate questioning as to whether, and how, attention should be devoted to it in the preparation of professional engineers.

The argument may be extended still further. The high standard of living which we enjoy and are striving to maintain and improve, and the establishment of a defensive strength which we hope we shall never have need to attempt to demonstrate, are dependent on our export trade. In this we are engaged in a technological battle with countries which can draw on richer natural resources and can bring larger numbers of trained men to bear on defined scientific and technical objectives. It is a battle in which national prestige is now linked with priority of scientific and technical achievement, and where what is achieved, or is said to have been achieved, in one part of the world—or well outside it—is reported everywhere within a few hours. Our scientists and technologists have become not only the guardians and guarantors of our national economy but also the agents of a new diplomacy in which more than an expanding export trade is at stake. They have a direct part to play as teachers or advisers in the industrialization of the underdeveloped countries and must not only carry conviction as experts in their particular scientific and technical specialities but also gain respect as ambassadors of broad vision, sound judgment and unquestionable integrity.

Again, it may be said that this latter is no new role for the Britisher—the point is that it is a role in which our scientists and technologists must play an increasingly energetic and responsible part.

It is into this complex and diversified activity that we senior members of our profession, in company with our colleagues i other branches of technology, are inviting a greater proportio of our most able young people. It is an activity which is worth of and needs the best. In so doing, however, we carry a heav share of the responsibility for ensuring that they are properl prepared by education, training and example for participatio in it. I shall try to analyse some aspects of what is involved.

In approaching this we shall be wise to avoid the view tha since we older men succeeded in some measure in adjustin ourselves without much systematic help to the rapidly changin situation I have outlined, the newcomers can well be left to d the same. Those who take this view and act accordingly ma be quite certain that many of the newcomers will deal with ne situations as competently as they themselves did in the pas But they would be wiser to recognize that the starting-poir has changed enormously, that the demands are growing in scop and severity, and that with increasing numbers there is likel to be a reduction in the average level of innate ability if nc in examination results. All these demand a more prolonge preparation in education and training. I do not mean by this however, an extended duration of undergraduate course (though this may prove to be desirable before it is physicall attainable) for reasons such as those expressed by Sir Richar Livingstone[1] in the remark

> If we wish to teach a subject, we must teach it at the age whe the mind can digest it. Otherwise we shall be like mothers wh feed their babies on beans and bacon. But there is another prin ciple, if not more important, even more commonly ignored. Th fruitfulness of education, at least in some subjects, depends o experience of life.

Engineering is one such subject. What I mean, therefore, i that the treatment of the principles of physical and engineerin science on which undergraduate and similar courses must largel concentrate will need to be supplemented increasingly at appro priate later stages of experience. This will demand the par ticipation as teachers of men whose primary activity is in th practice of engineering, and a much greater measure of collabora tion between the universities, technical colleges and industr than exists at present. Our limited present progress in thi connection, and other weaknesses of our educational system reflect our continuing tendency to regard the schools, th universities, the technical colleges and industry as separate com partments, and not as interwoven and interdependent elements in a continuous educational process. We in this Institution have done a good deal and must do much more to change thi situation.

Most of the preceding and of what I shall say later relate t the making of professional engineers. Unfortunately this questio is liable to obscure the importance of the supporting contribution and the needs in further education and training, of the muc larger body of our technicians and craftsmen. It is to frequently overlooked that without their skill and resource n new idea can be translated from the laboratory and design offic into efficient, reliable and economic engineering equipment. Th satisfactory education and training of these men—and women— is no less vital to the successful operation of industry and th public services, and to the health of the national economy, tha that of scientists and technologists, and in some respects it raise more difficult problems. I regret that space will not permit m to deal with them in this address.

THE SCHOOLS

Table 1 shows the response of the secondary grammar an public schools to the call of recent years that a larger number o our most able young people should prepare themselves for a

Table 1

GROWTH SINCE 1938 OF THE NUMBERS OF SCHOOL BOYS AND GIRLS PASSING THE ADVANCED LEVEL G.C.E. EXAMINATIONS IN PHYSICS, CHEMISTRY AND MATHEMATICS*

Year	Physics			Chemistry			Mathematics†		
	Boys	Girls	Total	Boys	Girls	Total	Boys	Girls	Total
H.S.C.									
1938	Not available		3153	Not available		3035	Not available		3224
1948			8589			7196			7508
G.C.E.									
1953	9654	1275	10929	8087	1454	9541	8185	1156	9341
1958	15382	2029	17411	11458	2185	13643	13541	1993	15534

* Supplied by the Ministry of Education.
† Excludes Applied Mathematics and Mechanics.

olonged study of science and its applications. With this call s been associated the growth of a closer contact and under- nding between the schools and industry—between those o are moulding the thinking of young people and those sponsible for providing opportunities for them to pursue sfying careers. This has occurred through vacation schools industry for masters and senior students; conferences repre- ntative of industry and the schools; the preparation of films icting scientific and technological progress; the issue of reer publications by the Press and of descriptive training oklets by individual firms; and the establishment in 1955 of e Industrial Fund for the Advancement of Scientific Education the Independent Schools.

If, in taking these steps, the aim on the industrial side has en to ensure that more of the best students devote themselves the study of science, there has nevertheless been explicitly ociated with it a desire to avoid a degree of concentration ch as might frustrate the achievement of a broad general cation. Certainly the Executive Committee of the Industrial nd had more in mind than a mere increase in the number of th-form science specialists. They were anxious that this reased number should be more broadly educated than has en the case. At age 18 our sixth-form scientists are probably ear ahead of their contemporaries anywhere in their knowledge science, but know correspondingly less about other subjects. e same applies, but even more so, to the arts specialists, for om only a modest attempt appears yet to have been made to ord an appreciation, limited though this may have to be, of history and achievements of science and of the manner in ich its progress has affected the evolution of our communal . I myself go a long way with Dr. James, Vice-Chancellor the University of Southampton,[2] in believing that

> he schools should endeavour to reach a situation in which the present design of studies in the Sixth Form is changed into one in which the child is not inevitably and finally committed to the arts or science side; in which education advances on a wider front, and he pupil's education, both in arts and science, goes on steadily, nd as far as possible, as one thing and not as two.

My knowledge of secondary school teaching is too limited me to be able to suggest how this objective might best be ieved. Much has been said and written by distinguished admasters during recent years. It has revealed the severity of conflict between the claims of general education, not least to mpensate for the cultural poverty of the home background many of the young people now participating in sixth-form rk, and those of specialist study, for the purpose of per- ting the more able to grapple with difficult concepts within ited field. But it does not require a deep knowledge of the schools to recognize that changes in desirable directions are being impeded by two factors, at least, for which the schools cannot be accounted responsible.

The first is the shortage of well-qualified science teachers, the scale of which may be judged from the data of Table 2.

Table 2

GRADUATE TEACHERS IN MATHEMATICS, PHYSICS AND CHEMISTRY IN MAINTAINED SECONDARY SCHOOLS PROVIDING G.C.E. 'O' LEVEL COURSES (OCTOBER, 1958)

	Mathematics		Physics		Chemistry	
	Men	Women	Men	Women	Men	Women
Total in posts, October, 1958	4461		2070		2375	
Net increase, 1957–58	93	81	123	1	134	38
	174		124		172	
Total vacancies and posts unsatisfactorily filled, October, 1958	511		217		111	

In comparison with the data of line 2 it is to be noted that, of the 1958 graduate (first degree) output in the above-mentioned subjects, 94 mathematicians, 177 physicists and 131 chemists accepted first appointments in the Government or public services or in industry, and that 118, 262 and 473 respectively elected to continue at the university for post-graduate work leading to higher degrees. Any expectation that a substantial proportion of the latter numbers may enter the schools in due course must unfortunately be restrained by the fact that, of those who left the universities with higher degrees in session 1957–58, only 2 mathematicians, 9 physicists and 15 chemists did so, compared with 14, 73 and 155 who joined the Government and public services and industry.

This is an untenable state of affairs and the consequences of a continuance of it will be disastrous. It is evident that in satisfying their need for more scientists the Government and public services and industry are incurring the grave risk of restricting the early scientific education of the greater numbers they are persuading to follow scientific and technological careers and of inhibiting its extension to all as an integral continuing part of general education.

The increasing magnitude of the problem may be judged from the data of Table 3, taken in relation to those of Table 1, which

show the Ministry of Education's estimates of the growth relative to 1959 in the number of 17-year-old students in the English and Welsh schools.

Table 3

ESTIMATED INCREASE RELATIVE TO 1959 IN THE NUMBER OF 17-YEAR-OLD STUDENTS IN SCHOOLS IN ENGLAND AND WALES

Year	1959	1960	1961	1962	1963	1964	1965	1966	1967
Relative number of students ..	100·0	118·0	129·5	148·0	139·5	178·0	200·0	182·0	179·0

The easy comment is that the salaries of graduate science teachers should be made more commensurate with those attainable in other forms of Government service and in industry. In times of shortage within a free society this is an elusive argument. In fact salary is by no means the only consideration affecting the willingness of graduate scientists to enter school teaching work. Of comparable importance is the improvement of teaching facilities and amenities, including the more generous provision of laboratory assistants, such as has been achieved in part with the aid of the Industrial Fund and as is now being extended by the Ministry of Education to the maintained schools. And professional men like ourselves can make a vital psychological contribution by seeking opportunities to give recognition to the great national importance of teaching work at all levels and by helping to raise the status of the teaching profession in the eyes of the general public. I return to the shortage of teachers in later Sections.

The second factor is that a broadening of sixth-form curricula can be made only with the collaboration of the universities, and there is all too little evidence of an active pursuit of such a collaborative effort. Regrettable though this may be, it is easy to understand. Such a change within the schools would involve a lowering of the present high standards of knowledge in specialist subjects required for university entry, which is in the wrong direction to assist the universities to withstand the pressure of new knowledge within these subjects. It would also increase the already considerable difficulties of selection. An increase in the duration of undergraduate courses from three to four years would be regarded as essential, only to be ruled out as unattainable in circumstances where the plans for university expansion will no more than cater for the increased numbe of new students to be expected from the data of Table 3.

The situation within the grammar and public schools is bes with difficulties and the changes which many regard as urger may be slow in maturing. Unfortunately there are evidences c movement within some grammar schools in the reverse directio I have in mind the introduction of engineering subjects into sixth form curricula, no doubt with the aim of stimulating the interes of some students in engineering as a career, and of providing a incentive to serious study. In my opinion it would be a grea misfortune if, through a misconception of the educational need of the professional engineer, the grammar schools were t encourage boys with little aptitude for mathematics and th basic sciences to remain at school to age 18 for the study o technical subjects. These boys should leave to enter industr at an earlier age and to pursue their studies at a technical college The profession of engineering is no haven for the scientifi weakling.

THE UNIVERSITIES

As a preliminary to consideration of the problems which fac the universities, it is necessary to review the recent expansio of their activities.

Table 4 shows the growth in the number of full-time student by faculties since 1938–39 and the fact that by 1958–59 th numbers of students of pure science had increased by a facto of 3·00 and of technology by one of 2·83, and that combine they now comprise about 38% of the overall university studen population. The expected further growth for which the Univer sity Grants Committee is seeking to make provision is to ar overall figure of 124 000 by the mid-1960's, of whom som 55 000 are expected to be students of science and technology, a compared with the present figure of about 38 000, with a large

Table 4

UNIVERSITY STUDENT NUMBERS

(supplied by the University Grants Committee)
Numbers of full-time students by faculties

Faculties	Academic years							
	1938–39		1948–49		1953–54		Autumn term 1958	
	Number	Per cent	Number	Per cent	Number	Per cent	Number	Per cent
Arts	22 374	44·7	37 147	44·4	34 673	43·0	42 739	42·9
Pure science	7 661	15·3	16 099	19·2	16 971	21·1	22 949	23·1
Medicine	11 883	23·8	13 718	16·4	13 239	16·4	12 688	12·8
Dentistry	1 488	3·0	2 547	3·0	2 564	3·2	2 980	3·0
Technology	5 288	10·6	10 884	13·0	10 036	12·4	14 942	15·0
Agriculture and forestry	1 043	2·1	2 919	3·5	2 066	2·6	2 032	2·0
Veterinary science	265	0·5	376	0·5	1 053	1·3	1 206	1·2
Total	50 002	100·0	83 690	100·0	80 602	100·0	99 536	100·0

Table 5

POST-GRADUATE UNIVERSITY STUDENTS
(supplied by the University Grants Committee)

	1938–39	1948–49	1953–54	Autumn term 1958
Science	1268	2429	3290	3853
Technology	388	858	1318	1916
Total in universities	3094	6452	9468*	11062*

* Total excludes students working for a Teacher's Diploma.

-oportionate increase in technology than in science. Table 5 ves the growth of post-graduate student numbers with par- cular reference to those in science and technology.

Undergraduate Courses

In the preceding Sections I have referred to the pressure on ndergraduate courses which is being generated, on the one and, by a growing recognition that the work of the sixth forms school—and also of the undergraduate courses themselves— too specialized, and on the other, by the extremely rapid pansion of knowledge in specialist fields, particularly within ience and technology. As already mentioned, the simple swer to both would be to increase the duration of under- aduate courses by a year, and some would find further justifica- on for such a change in the greater length of the corresponding urses in Continental countries. For the majority of students, owever, most of all for those of technology, three years at a niversity is a long enough preliminary to active participation in e affairs of life outside. Of greater urgency is the need for the niversities to rethink the objectives and content of their present ree-year courses, and for them to consider co-operatively nong themselves, and in consultation with the technical colleges d industry, their distinctive contribution to post-graduate udies for a proportion of the graduate output. This is made rticularly necessary in the field of technology by the develop- ent within the technical colleges of courses leading to the new iploma in Technology and by the recent announcement of a gher award, that of Membership of the College of Tech- logists, to which I refer in the next Section.

I appreciate in respect of undergraduate courses in electrical gineering that much has already been done to remove or reduce tention to many of the technicalities which used to find a ace in them. My doubt concerns whether enough recognition s yet been given in these courses to the dependence of progress electrical engineering on the work of physicists, chemists, athematicians and metallurgists, and to the outstanding portance of the development and exploitation of new materials. e teaching of the subject of electrical machines, and of electro- chanical systems generally, is particularly in need of recon- deration with a view to making it as stimulating and searching study as that of electronics and communications and to rrecting the unbalance of student interest in favour of the ter which has prevailed for some time. Still better, to show w the two are interrelated through the principles of electro- agnetism and the properties of materials, and bound together a common mathematics. Considerable initiative is being ken in this matter within the Electrical Engineering Depart- nt of the Massachusetts Institute of Technology, where the w is held that the analysis of electro-mechanical systems in ms of steady-state concepts using equivalent circuits and twork theory does not afford a sufficiently progressive proach to the problems of electromagnetic energy conversion, and that there may be merit in starting from the complete equations of the most general system derived from Hamilton's principle, energy considerations and field theory and in teaching the student to deal with particular cases by the introduction of appropriate simplifications into the general case.

I do not underestimate the difficulties—the best method of approach will not be easy to find. I know it will be said, from personal experience of similar attempts, that undergraduate courses which drew as extensively on science and mathematics as I am suggesting would be unattractive to students who, having chosen engineering rather than science, expect their course to introduce them to the practice of their chosen subject and to permit some concentration on a specialist branch of it in the final stages. What I visualize is no contradiction of this expectation. The problem is how to broaden the fundamentals without making, or appearing to make, these fundamentals an end in themselves; and how therefore to illustrate their technological significance by examples drawn from developing practice as a bridge between scientific abstraction and real engineering.

The desirable result will not be achieved merely by the incorporation of specific courses in physics, chemistry, mathematics and metallurgy, taught, in a manner apart, in the sense of the honours schools in these subjects. Nor, of course, must the treatment of their appropriate parts carry with it the stigma which I have known to be associated with courses designated 'for engineers' mainly because the scientists and mathematicians concerned were inclined to regard them as a provision for second-rate minds rather than for first-rate ones whose interests lay in the application of these subjects, as distinct from the pursuit of them for their own sake. The attitude of intellectual superiority which has too often characterized the contribution of scientists and mathematicians to engineering education has been a major factor in delaying the emergence of engineering to the status it is now coming to occupy as a subject of university study, and as a pursuit worthy of, and needing, a substantial proportion of the best brains available. The trend of university education in electrical engineering which I believe to be desirable will demand a fuller measure of inter-departmental collaboration and a less rigid adherence to traditional departmentalism within science and technology than has existed in the past, and than appears still to characterize too many of our universities.

Even then the full potentialities of a university education will not have been realized unless the student's mind has been opened to an appreciation of the fact that exclusive devotion to a scientific discipline will not give him all he needs as a preparation for life. With this as his theme, Sir Eric Ashby has presented the case for an extension of inter-departmental collaboration beyond science and technology into the domain of the arts faculties in a stimulating paper entitled 'Technological Humanism'.[3] In this he says,

> If technologists are to enter fully into their responsibilities in the world of tomorrow, they have to adapt themselves to novel technologies; they must also be able to adapt themselves to the social consequences of the novel technologies. To pursue technology in a spirit of liberality, they need not only a thorough training in the theory and practice of technology; they need also an understanding of men and communities such as is acquired through literature and history and the social sciences.

And he gives examples of the type of course which he considers should be made available to students of technology by the faculties of arts.

Some universities have already taken steps in this direction, and I hope others will follow. I am naturally particularly aware, for example, of the scheme of lunch-time lectures at the Imperial College of Science and Technology, and of the opportunity available to the final-year undergraduates of this College to spend a day per week at the London School of Economics. In

my opinion, however, experiments such as these will be really successful only if the teachers of technology provide a bridge, by themselves constantly referring to and illustrating the humanitarian aspects of technology in the presentation of their respective subjects. And it should not be overlooked that perhaps the most liberalizing influence of all is active participation by the student in the corporate life of the university.

Post-Graduate Courses and Research

Movements in undergraduate engineering education in these directions will, of course, leave for later acquisition an increasing amount of more specialized knowledge. With varying degrees of relevance to different men, this concerns the advanced aspects of the engineering science of their chosen field, the techniques of the various specialized branches of it, the materials on which the successful exploitation of these techniques are based, and the use of the analytical tools which are now available for the solution of complex engineering problems.

Until quite recently the acquisition of this supplementary knowledge was left largely to individual effort and initiative, and little provision was made in the form of post-graduate courses for its systematic presentation. However, several university engineering departments are now providing courses of one year's duration on subjects related to their particular research interests; the larger technical colleges are organizing part-time, mainly evening, courses in subjects of special relevance to the industries of their respective regions; and in some industrial concerns lecture courses and discussions on technical and organization matters form an integral part of their graduate training schemes. It is beyond question that the need for such courses will grow and that participation in them must be recognized and accepted as an increasing necessity.

In respect of the university provisions, I hold the view that participation in them should desirably be preceded by a period of industrial training as a foundation for the reliable choice of a specialist field, and so that the graduate may gain an appreciation of the grounds on which existing practices in his chosen field have emerged and of the directions in which improvement is required. I am aware that many of the university staff members concerned also hold this view; that they are disappointed with the response of industry to the steps they have taken; and have concluded that the only means of ensuring adequate support for their courses is to accept as students men who have just graduated.

The limited response of industry has been due in part to the shortage of graduates at a time of rapidly expanding industrial activity, and I hope that the situation will improve with the increasing annual output of graduates and with the termination of the National Service arrangements. But in part the universities are themselves responsible.

In the first place it would have been wise, in the circumstances just mentioned, had they not all adhered so rigidly to the normal sessional pattern of university courses, but had been willing to offer, as an initial step, courses of shorter duration. Secondly, some of the courses offered have been essentially departmental in character, have not drawn on the wider resources of their university, and have therefore failed to reflect the kind of interrelationships within science and technology to which I referred earlier. And thirdly, there has been some reluctance to recognize the need at this level for the participation of outside experts as lecturers. In this respect I believe we have something to learn from other countries in which, for limited periods, carefully selected men whose primary employment is in industry or Government establishment are given academic status and contribute to the advanced specialist teaching on a part-time basis. It is quite unrealistic to suppose that the teaching of technology at its highest level can be carried wholly, or largely, by men who occupy full-time teaching posts, unless these men return to industry or a similar environment at appropriate intervals. In my opinion the raising of technological teaching in this country to the level required demands a much greater mobility of staff between educational institutions on the one hand and industry or Government establishment on the other than has yet been achieved, and, in supplementation, that carefully selected men should be fully recognized members of both, as is already the case in medicine. A good start would be for the research establishments of the Department of Scientific and Industrial Research, and some of the research associations, to be accepted by the appropriate universities as recognized institutions in this sense. The separation of the D.S.I.R. establishments from the domain of technological teaching, and the lack of a recognized link between them and the universities in respect of scientific and technological research, are national handicaps of great seriousness which should be removed as quickly as possible.

The universities have experienced considerable difficulties in recent years in obtaining as full-time staff members of their engineering departments a sufficient number of men who are leaders, or potential leaders, of thought in their subject and also keen and capable of teaching it. The difficulty arises basically from the fact that, in deciding to take up the study of engineering, a young man is motivated, generally speaking, by the desire to practise, not to teach, it. His entry into teaching work usually involves, therefore, a withdrawal from practice under circumstances where, at present, the practice affords considerable attractions. Salary is perhaps most frequently quoted in this respect, but while it would be idle to deny the significance of this in some cases, it is often overstated in the light of recent salary changes, and of the possibilities, for the best men, of outside consulting work. Of greater moment may be the limited flexibility of industrial superannuation schemes in respect of the transferability of pension rights. But the difficulty goes deeper. For men with engineering in their blood, the practice of it, particularly at times of rapid technological change and industrial expansion, is a highly stimulating experience and one which affords considerable scope for the exercise of initiative and the carrying of responsibility. The problem is how to marry this situation to the need, in the larger technical colleges as well as in the universities, for more forward-looking teachers of technology.

In general, first-class men will not be attracted, nor will they remain first-class and maintain a vital environment for their students, unless they are afforded the opportunity of pursuing research work. It is therefore gratifying that the Department of Scientific and Industrial Research has now been provided with augmented funds for the support of research programmes of special timeliness and promise. The dangers are that the pursuit of research in specialist fields may become a dominant interest, at the expense of attention to the broad educational needs of the undergraduate student body and of adequate personal acquaintance with the individual members of this body; and that the attempt to synthesize new knowledge and to unify it with the old for teaching purposes may be neglected because it does not carry the same glamour nor gain the same recognition as research (as normally understood). Moreover, the retention of raw graduates, for participation in limited research programmes, tends to generate in these young men an inability to recognize on their entry into industry that research work is not a thing apart, but something which, to be effective, must be woven into the whole fabric of industrial activity; and that it requires of the researcher experiences which cannot be gained in isolation from this wider environment. It is this

kind of consideration which prompted the conception of the Membership of the College of Technologists award to which I refer in the next Section.

I recognize that much has been done within the universities towards attainment of the objectives I have tried to develop in this Section. I also recognize that it is a good thing for there to be differences between universities, even substantial ones. But there seems to me to be great need for systematic inter-university discussion at faculty level, not only in the direct interests of the universities themselves, but also because collaboration with industry and Government establishments would be facilitated and strengthened thereby, and because they would be the better able, as in the national interest I believe they should, to assist the colleges of advanced technology to formulate the distinctive contribution to technological education which it is hoped they will make. A clarification of the position which these colleges are to occupy in the structure of technological education may soon become a matter of urgency.

THE TECHNICAL COLLEGES

Prior to the late war the only means open to the vast majority of young people who entered industry direct from school and who wished to acquire the knowledge and qualifications which they saw to be necessary for their progress was attendance at evening classes. These classes remain a very important, and indeed a growing, part of further education, but by themselves they are no longer adequate to provide for the technical needs of developing industry, let alone for the needs of the individual for continued general education. There has been a gratifying increase since the war in the willingness of industry, and of the electrical industry in particular, to release its young employees to attend part-time day courses. Their number increased from 41 460 in 1938 to 434 672 in 1958, but very wide discrepancies exist between different industries in the granting of this facility, and there is much scope for further improvement. In its White Paper on Technical Education (February, 1956), the Government set 700 000 as its target for the number of part-time day students by the early 1960's, and it is to be hoped—though there is reason to doubt—that this figure will be achieved.

Table 6 shows the scale of development of Ordinary and Higher National Certificate courses since 1938. These courses are now largely operated on a day-time basis, and this has reduced the wastage previously associated with evening study. But the wastage is still very large. The basic reason is that many of the boys admitted to the early years are unsuited to this type of course. They enter with their eye on the fact, as shown in Table 7, that National Certificate courses have been a major

Table 6

GROWTH SINCE 1938 OF STUDENTS QUALIFYING FOR ORDINARY AND HIGHER NATIONAL CERTIFICATES*

Subject	1938		1958	
	O.N.C.	H.N.C.	O.N.C.	H.N.C.
Electrical engineering	917	379	4 271	2 165
Mechanical engineering	1 449	502	8 599	3 926
Production engineering	—	—	—	542
Civil engineering ..	—	—	—	278
Building	544	137	1 597	929
Chemistry	151	62	1 437	852
Applied chemistry ..	—	—	22	30
Chemical engineering	—	—	—	61
Applied physics ..	—	—	405	169
Metallurgy	—	—	373	269
Naval architecture ..	16	—	100	47
Textiles	104	57	129	49
Mining	—	—	715	205
Mining surveying ..	—	—	—	118
Commerce	132	—	380	7
Totals	3 313	1 137	18 028	9 647

* Supplied by the Ministry of Education.

Table 7

ANALYSIS OF MEANS OF COMPLIANCE WITH THE EXAMINATION REGULATIONS OF THE INSTITUTION 1954–58 INCLUSIVE

Calendar year	Degree	Diploma or Associateship	Via National Certificate	Institution Examination	Miscellaneous	Total
Elections and Transfers to Graduate Membership						
1954	469	179	707	126	6	1 487
1955	448	144	770	59	2	1 423
1956	580	179	1 101	81	9	1 950
1957	627	171	1 413	26	3	2 240
1958	587	158	947	19	3	1 714
Direct Elections to Associate Membership						
1954	103	12	87	24	15	241
1955	80	14	68	18	20	200
1956	135	19	115	28	23	320
1957	128	13	111	23	27	302
1958	130	22	112	8	24	296

route to exemption from the examination requirements of the professional institutions, and are inclined to attach little value to other courses, much more suited to many of them, which lead to technician and craft qualifications of the City and Guilds of London Institute and similar examining bodies. So much so that, having failed, too many withdraw completely from further study. No doubt the situation will be improved by the clarification of the structure, and of the interrelationships, of courses at the preliminary stage which is now occurring. But basically its solution demands of industry a much clearer recognition of the value which attaches to success in technician and craft courses, and closer collaboration with the technical colleges in the encouragement of its apprentices to participate in them.

The normal duration of day-time release from employment is one day a week for about 36 weeks per year. On this basis, with some evening work at college added, satisfaction of the examination requirements of the various professional bodies is becoming an increasingly difficult undertaking. With our own Institution it now involves a minimum of eight years from entry to the first year of the Ordinary course. Even then the sequence of scientific and technical subjects is not best suited to professional purposes. Moreover, devotion of the whole period of college attendance to scientific and technical study affords no opportunity for participation in the corporate life of the college, and no time for the continuance of general education except in so far as good teachers may achieve something of this within the scientific and technical subjects themselves.

These limitations have been overcome by the sandwich arrangements which characterize the new Diploma in Technology courses administered by the National Council for Technological Awards. Their normal structure is of alternate periods of six months in industry and in college over a four-year period, with entry based on an average mark of 60-70% in the Ordinary National Certificate examination, or Advanced Level G.C.E.

passes in mathematics and physics or chemistry, plus four passes at the Ordinary G.C.E. level.

Table 8 gives the number of students who were pursuing Diploma in Technology courses during the 1958–59 session.

Table 8

NUMBERS OF DIP.TECH.(ENG.) AND DIP.TECH. STUDENTS SESSION 1958–59

Subject	Number of students				
	First year	Second year	Third year	Fourth year	Total
Dip.Tech.(Eng.):					
Electrical engineering	358	299	143	67	867
Mechanical and production engineering	361	274	127	40	802
Civil engineering	23	3	2	—	28
Chemical engineering	37	22	2	—	61
Instrument and control engineering	10	9	—	—	19
Aeronautical engineering	64	22	—	—	86
Building	8	—	—	—	8
Dip.Tech.:					
Applied biology	9	6	—	—	15
Applied biochemistry	5	—	—	—	5
Applied pharmacology	5	—	—	—	5
Applied chemistry / Chemical technology / Industrial chemistry	103	58	34	9	204
Mathematics	40	14	—	—	54
Metallurgy	32	34	26	6	98
Physics	116	88	47	15	266
	1171	829	381	137	2518

Of these, some 35% had entered from Ordinary National Certificate courses. Success in the courses in electrical engineering so far approved by the National Council will afford complete exemption from the examination requirements of The Institution, and it is to be hoped that this will apply to all further courses which are similarly approved. As indicated in a joint statement[4] by the Ministry of Education and The Institution, published in 1958, it is likely that there will also be an increasing provision of similarly organized sandwich courses with a somewhat lower standard of entry and of attainment leading to Higher National Diplomas and partial exemption from The Institution's examination requirements.

These developments, in association with the trend for academically promising boys to remain at school beyond the statutory leaving age, suggest that in due course the great majority of prospective professional electrical engineers will either not enter National Certificate courses or will leave them at the Ordinary Certificate stage, and that these courses will come to serve mainly the needs of high-grade technicians, as was the conception when they were introduced some 30 years ago. The Institution has no intention, however, of closing this route to professional recognition. There are always likely to be some men of professional ability for whom, because of their inclinations or domestic circumstances, or because they are unable to arrange with their employers for their participation in a sandwich course, this will be the only route. The Institution has been criticized for too rapid an introduction of its new examination regulations to the embarrassment, reflected in the 1958 figures of Table 7, of many men now engaged on National Certificate courses, and I am sure it will be the wish of all of us to minimize such hardship as may arise during the transitional period.

When inaugurating the Diploma in Technology scheme the National Council for Technological Awards declared its intention to establish for the Dip.Tech and Dip.Tech.(Eng.) awards a standard comparable with that of an honours degree. This comparison inevitably raised the question as to what were to be the distinctive features of diploma in technology and degree courses. It was obvious that they were structurally different, and the significance of this was emphasized in the National Council's statement that the periods of industrial training were to be regarded as just as much a part of the course as those of college study, and that the two parts should be closely co-ordinated. An important factor in this concept was that in general the students would be 'industry based'; that is, they would be attached to sponsoring industrial concerns as student apprentices and these concerns would therefore act in partnership with the college in the achievement of this co-ordination. I shall refer to this matter in the next Section. A second distinction was the National Council's insistence on the inclusion of liberal studies within the curricula. What has not yet emerged explicitly is whether there are to be two distinctive approaches to the teaching of the technological subjects themselves; whether in fact there can or should be for men aiming at the same professional objective. It is easy to say, with the sandwich character of the diploma courses in mind, that the presentation in these courses should be more closely related to practice in the subject of which the student is gaining interwoven industrial or allied experience. The difficulty lies in defining what this means without contradicting the necessity, which I have already elaborated, for the teaching to prepare the student for a future, not merely for the present, situation. The broad distinction in method I myself would draw is that the diploma courses should tend to work from the particular towards the general, and degree courses from the general to the particular. Changes in university undergraduate courses in the directions discussed in the preceding Section, while the diploma courses are establishing themselves on their present lines, would contribute considerably to the drawing of this distinction. I do not see, at the moment, why the two approaches should not provide comparable intellectual disciplines. The matter affords much scope for educational research of which far too little is being done within, or if done, being published by, either the universities or the technical colleges. A number of well thought out papers on it would be a welcome and valuable addition to the literature of The Institution.

I remarked previously on the need for supplementation of our 3-year undergraduate courses by systematic post-graduate studies. The same arguments apply to the 4-year Dip.Tech and Dip.Tech. (Eng.) courses, which many feel should soon be extended to 5 years. In this case, accommodation, within the colleges of advanced technology at any rate, is unlikely to be the restraining factor. More important considerations are likely to be staffing and the possibility that firms which are at present financing students throughout a 4-year course may be unwilling to give them longer support. However, whether linked directly to an extension of Dip.Tech. courses or not, it seems evident that the larger technical colleges must come to organize full-time advanced courses of appropriate duration in the specialist fields of technology relevant to their respective regions, in supplementation of the advanced evening courses they are already providing.

Developments along these lines will have a special significance to the recent decision of the National Council for Technological Awards to institute an award higher than the Diploma in Technology, that of Membership of the College of Technologists. This award will normally be made for a programme of work successfully carried out jointly in industry and at a college, and

registration for it will be available to university graduates as well as to holders of the Dip.Tech. or Dip.Tech.(Eng.). The programme may be concerned with any technological aspect of industrial activity—for example, research, development, design, production or market investigation—and submission for the award may be made three years after approval of the project and registration as a candidate, provided a high proportion of the candidate's work in industry has been devoted to the project and this work has been adequately supported by college study. A particularly important feature of the scheme is that the candidate's work will be supervised jointly by staff members of the industrial organization concerned and of the college at which the advanced study, or, where appropriate, some of the research, is to be undertaken. This serves to emphasize the point made in the announcement of the scheme that 'the integration of academic study and industrial training is an important feature of courses leading to the Diploma in Technology, and the Council feel that a similar collaboration between industry and college should be extended into the post-diploma field'. I would add that this is another field in which there is need for closer collaboration between the larger technical colleges and the universities if most effective use of staff and facilities is to be ensured.

In the preceding Section I referred to the staffing of the university departments of technology. The needs of the technical colleges in this respect were reported upon in 1956 by a Special Committee of the Ministry of Education,[5] and Table 9 is reproduced from the report.

Table 9

ESTIMATED NET ANNUAL INCREASE AND GROSS ANNUAL RECRUITMENT OF FULL-TIME TEACHERS IN TECHNICAL COLLEGES REQUIRED OVER THE FIVE YEARS 1956–61

	Total number employed 1955–56	Total number of extra teachers required by 1960–61	Annual averages			
			1952–55		1956–61	
			Net increase	Gross recruitment	Net increase required	Gross recruitment required
All full-time teachers	11 455	7 160	770	1 300	1 430	2 300
Teachers of professional courses in science and technology:						
Mathematics and science graduates	(2 250)*	1 260	200	300	250	400
Technology graduates	(1 250)*	1 325	90	150	265	360
Graduate-equivalent	(900)*	190	75	120	40	100

* Not accurately known, but estimated from Ministry statistics for 1954–55.

The latest figures available show that during the 1957–58 session there was a net increase of about 1 500 in the overall number of full-time teachers as compared with the estimated average annual requirement of 1 430; that the estimated annual gain of 250 mathematics and science graduates was just met; but that the number of graduate teachers of technology increased by only 150 as against the requirement of 265. Since the latter figure was also not met in 1956–57, there is an accumulating deficiency in respect of this category of teacher, and the position at the senior lecturer and reader levels gives particular cause for anxiety.

There are three other matters which I feel I must mention. One is the gratifying increase which is occurring in the number of suitably qualified candidates for entry to the technical teacher training colleges as a preliminary to technical college appointments, and the recent decision of the Ministry of Education to expand the facilities of these colleges considerably. This augurs well for an improvement in the standard of technical college teaching of which the universities might wisely take note. The second is the prospect of the early establishment of a residential staff college in which technical college teachers and representatives of industry, Government establishments, universities and the schools could examine and discuss together the aims and methods of technical and technological education in the light of experiences and developments in these related fields. The third concerns the need for the lively participation of senior industrial representatives in the government of technical colleges. Strong governing bodies and active advisory committees are vital to the development of a closer partnership between them and industry and to the necessary improvement in their status and staffing.

THE TRAINING OF PROFESSIONAL ELECTRICAL ENGINEERS

The preceding three Sections have dealt with certain aspects of the contributions of the schools, universities and technical colleges to the education of technologists. The major Engineering Institutions have always held the view, and have recorded it in their bye-laws, that for the adequate preparation of professional engineers this education requires to be supplemented by a further process, normally conducted in the environment of engineering practice and referred to as 'training'. The need to use this distinguishing word in relation to prospective professional engineers is perhaps unfortunate, since it is responsible for a widespread misunderstanding abroad of the essentially educational objectives involved. In respect of our own branch of the profession, these objectives are broadly to ensure:

(i) A sound knowledge of the properties, uses, limitations and methods of handling of the conducting, insulating, magnetic and structural materials employed in the construction of electrical equipment.

(ii) An appreciation of the nature and interrelationships of the main functional aspects—research, development, design, manufacture, sales, installation and operation—of electrical engineering activity, and of the respective contributions which need to be made to them by professional engineers, technicians and craftsmen.

(iii) A comprehensive acquaintance with the various facets of a selected specialist field and of other related ones on which the practice in it is dependent.

(iv) An understanding of the reasons, administrative and economic as well as technical, for present practices, and an enthusiasm to search for the directions in which improvements might most profitably be pursued.

Over many years The Institution has had Examination Regulations governing its educational requirements for Graduate Membership. These are about to be supplemented by Training Regulations, and candidates for Associate Membership will be required to submit a certified statement detailing the manner in which they claim to have satisfied these regulations. It is expected that this statement will often be submitted by Graduate Members some time prior to their application for Associate Membership with a view to obtaining an evaluation of their training, and, when completed, registration of the fact that it has been accepted as satisfactory. In respect of university students the normal procedure for meeting the requirement will be by post-graduate training of two years' overall duration, inclusive, where appropriate, of training obtained during long vacations and of periods spent in attendance at courses of advanced study. In some cases 9–10 months of the 2-year period will have been completed between leaving school and entering the university. I myself wish that this 1 : 3 : 1, or thick sandwich, arrangement could be made more general. For holders of the Dip.Tech.(Eng.) it should be met by the training they will have received during the industrial periods of the sandwich course, supplemented by the final phase of training normally provided for in student apprenticeships. While the position of those who have satisfied the Examination Regulations wholly by part-time study cannot be stated so clearly, the overall duration of their apprenticeships will usually have been much longer than for either of the previous categories, though the pattern and content may sometimes not have been ideal for professional purposes.

In appropriate cases the Council propose to accept, in lieu of formal training, work carried out in a staff appointment under the supervision of a Corporate Member. This will involve the submission of a certified statement detailing the character of the work and the range of basic experience it has afforded. If a part or the whole of the work cannot be accepted as training, it may be treated as 'additional experience' and as such will compensate for a deficiency in training on the basis that *two* years of 'additional experience' will compensate for *one* year's deficiency in training. In making their assessment the Council will take special note of any post-graduate course of an appropriate nature undertaken during this period.

For many years past the larger, and some comparatively small, firms of the electrical industry, and also the operating organizations, have devoted a great deal of care and much of their resources to the planning of training schemes for achieving the objectives to which I have referred, and have associated with the systematic movement of their graduates and student apprentices from department to department participation in lecture courses and discussions on the technical and organizational features of their concerns. These are not the only concerns, however, which need the services of men who have received a broad and thorough training. So far, the others have managed, when particular needs arose, by recruiting well-trained men from elsewhere. While in the interests of individual advancement no one should question the desirability of reasonable mobility from firm to firm, the scale on which movement has occurred during recent years has been highly inequitable and extremely wasteful. A more important consideration for the future, however, is the fact that the responsibility for training the increasingly large national pool of technologists—and, no less important, of technicians and craftsmen—required to accelerate the technical changes and achieve the enhanced productivity which are now possible cannot be carried by the few firms which have had the foresight to provide training facilities. Moreover, additional comprehensive training facilities must be made available as a matter of urgency on behalf of the increasing output of university graduates and if the Dip.Tech. scheme is to achieve the standards and objectives for which it was established. The shortage of these facilities has indeed become the Achilles' heel of our national plans for the further development of technological and technical education.

It would seem that this problem can be resolved only if the smaller and more specialized firms will collaborate in the organization of group schemes in which their limited individual resources are properly co-ordinated. A good example of such a scheme for the training of graduates and of student apprentices is the Scottish Electrical Training Scheme embracing five electrical manufacturing concerns and two Electricity Boards. Unfortunately there is not much evidence of initiative elsewhere to follow this example. It is a matter of speculation, therefore, as to how an extension of it is to be achieved, and, regrettably, of doubt as to whether it can ever be achieved on a sufficient scale in the absence of some form of financial incentive provided by the Government. The National Council for Technological Awards has appealed to the firms providing training for Dip.Tech. students to assist in ensuring that the content of the college and industrial parts of the sandwich courses are closely co-ordinated. But, as compared with the care it has taken in assessing the college facilities, their staffs, and the structure and curricula of the courses, and with its requirement for the appointment of external examiners, it has not yet been able to make a similar assessment of the nature and standards of the associated industrial training. This it is now approaching through the medium of a training panel, and no doubt the new Training Regulations of The Institution will be of assistance to it in this delicate and difficult matter.

It would be less than fair, and indeed incorrect, if having commented rather critically on certain aspects of technological education, I were to appear to profess that all is well with the training provided by the large electrical concerns to which I previously referred so favourably, but of which I know some of my late university colleagues to be in their turn far from uncritical. They remark that the training arrangements tend to be too inflexible to cater adequately for the more intelligent graduates; that too often the part spent in the workshops fails to avoid boredom and frustration; and that the opportunities afforded for the exercise of initiative and for the carrying of preliminary responsibility are too long delayed. These are very real problems. Their satisfactory resolution involves not only the provision of internal lecture courses and discussions on technical, organizational and economic aspects of the firms' activities, but also the planning of well-conceived projects to be carried out by individuals or groups under appropriate guidance and supervision. Though a good deal has been attempted in these directions in some cases, it cannot be questioned that there is considerable scope for further experimentation in pursuit of the aims to which I referred at the beginning of this Section. It is important to recognize, however, that this is a matter to which industrial training officers can make only a limited, though vital, contribution. They must receive the enthusiastic support and active help of men occupying positions of authority at all levels of their organizations, and particularly of those who are Corporate Members of The Institution. It has been a surprise and disappointment to me to observe how little some graduates are prepared to do to assist those who are seeking for information of which they themselves must have felt a similar need only a few years previously.

Even so there are limits to the help which should, or could in any case, be guaranteed. A real urge is stimulated, not inhibited, by difficulty, and discouraging circumstances will seldom fail to provide opportunity for those who are determined to find it. Those who complain most of the lack of opportunity

are usually those least capable of recognizing and accepting it. The taking of responsibility may be impeded, but, as evidenced by the careers of many of our predecessors, cannot be denied to those who are determined to have it.

I am aware also that some senior members of industry, more particularly on the electronics side, are not persuaded as to the value of organized schemes of graduate training, and consider that all necessary experience can be obtained in a staff appointment. As already mentioned, this has been provided for within the new Training Regulations. I think it relevant to mention, however, that the diploma in engineering of Continental countries is seldom awarded on less than 5 years of college study from age 18, and that its award usually requires up to 12 months of industrial 'training', of which, in some cases, 6 months must be obtained before entry upon the diploma course. It would seem to me a risky policy, therefore, to assume that a 3-year course of engineering science in this country affords by itself an equivalent preparation for responsible employment.

MANAGEMENT STUDIES

It is a big step from satisfying the examination requirements of one of the professional institutions to the attainment of senior executive positions on the research, development, design, manufacturing, commercial or operating sides of engineering work, for this involves much more than profitable participation in a scheme of training and in advanced courses such as I have mentioned. The ability to acquire knowledge is one thing—the ability to carry responsibility and exercise authority in relation to this knowledge is quite another. The only effective means of establishing that a man can carry responsibility is, of course, to put him in a position where he is obliged to do so. In no other way can he demonstrate the possession of certain essential personal qualities and abilities. When this has been demonstrated by success in one situation, it is a characteristic of good management that he will be tested and matured in others of greater complexity, with the advice and example of his more senior colleagues to guide him.

Nevertheless, a great deal of thought has been given during recent years to the question of what can be done outside the immediate environment of employment to assist this process. Is there a framework of general principles and a systematic body of knowledge about industrial organization and management which, properly presented, will guide and assist young engineers the better to assess their personal experiences in this field? Can anything be done—in courses—to develop a man's ability to recognize and appreciate the significance to his own field of work of the activities of others engaged in related fields; to organize co-operative effort among men of whose respective specialities he may have only limited detailed knowledge; to draw reliable deductions from, and make sound decisions on, inadequate evidence and in insufficient time; and to represent the views of other people, as well as to express his own, clearly and concisely?

Much has been written on these, and other related, questions and many post-graduate experiments are being tried within the universities and technical colleges, and through the medium of staff courses organized by industrial concerns and public authorities. The anxiety attaching to the former, particularly where the students lack outside experience, is lest they may assume that successful participation in a course of study will automatically endow them with the qualities essential to good managers. It is of the greatest importance that this danger should be avoided.

THE UTILIZATION OF PROFESSIONAL ENGINEERS

So far, I have concentrated on the aims of engineering education and training and on some of the problems which are associated with the early stages of their achievement in circumstances where the student body and the facilities for accommodating and teaching it are expanding rapidly. The scale on which new facilities have been and are being provided has been governed by the abnormally high birth rate in the mid-1940's and by the need to correct the shortage of professional scientists and engineers to which frequent reference has been made during the past few years. A very important factor affecting our thinking on the latter point has been the comparison drawn between the output of similarly qualified scientists and engineers in this and other industrial countries. In his recent Rede Lecture in Cambridge, Sir Charles Snow[6] expresses the position as follows:

> Our population is small by the side of either the U.S.A. or the U.S.S.R. Roughly, if we compare like with like, and put scientists and engineers together, we are training at a professional level per head of the population one Englishman to every one and a half Americans to every two and a half Russians. Someone is wrong.

Taking this statement literally, it means that the U.S.A. is producing annually $4\frac{1}{2}$ times, and the U.S.S.R. 10 times, as many as we are.

Although these figures may paint rather too pessimistic a picture, and do not of course depict the relative starting-points in accumulated present numbers, they nevertheless give an indication of what is happening elsewhere. Of more direct relevance, however, are the results of an attempt by the Advisory Council on Scientific Policy in 1955–56 to quantify our own needs by an extensive inquiry among employers. These results were published in a White Paper entitled 'Scientific and Engineering Manpower in Great Britain', which recommended that the annual output of scientists and engineers of professional standard should be increased from the 1956 figure of around 10 000 to some 20 000 by the late 1960's. There seems no reason to doubt that this target will be met.

In my opinion we should have been culpable had we aimed at any lower figure. It is therefore with no intention of questioning it that I ask, Are we using the highly qualified persons we already possess to the best national advantage, and, if not, what steps should we take to correct the situation?

In an article in the March, 1959, *Journal* of The Institution entitled 'Recruitment of Professional Electrical Engineers', G. S. C. Lucas makes an analysis of the growth of our corporate membership during recent years and an attempt to project it into the late 1960's. In doing so, he suggests that in the course of their professional work some of our Corporate Members may be performing duties which, with appropriate replanning and with the fuller utilization of computing and other new techniques, might well be undertaken by men or women of lower academic qualifications than are required for corporate membership. This process of devolution is, of course, a continuous one in progressive industry, and it is characteristic that highly qualified people so relieved move into new fields of further creative effort with similar later consequences. Perhaps surprisingly at first sight, it is a process which increases rather than decreases the need for highly qualified people. But it demands for its progression the availability of people, themselves well educated and trained but to a different level, upon whom the work can be devolved. I believe in this connection that in stressing our shortage of scientists and technologists we have given inadequate recognition to the importance of the supporting contribution, and too little attention to the educational needs in the rapidly changing situation, of the much larger body of our technicians and craftsmen. Unfortunately, I cannot

develop here this aspect of our professional responsibility, but I have discussed it elsewhere,[7] and I would only add that I think it of great importance that we should help to devise a means of affording appropriate status to the high-grade electrical technicians on which we professional engineers are so dependent. Unfortunately, although a great deal of thought was given to this matter within The Institution a few years ago, no satisfactory solution was then forthcoming. I hope it may receive early reconsideration, since a solution would do much to allay the criticism which has been levelled at the new Examination Regulations.

There is a further aspect of our utilization of scientific and engineering manpower about which I find myself becoming increasingly concerned. It arises from the high degree of subdivision which characterizes our industrial structure. This subdivision produces among the larger organizations a considerable multiplication of research, development and design effort directed to the same, or substantially the same, technical objectives, and the possibility that no one of these concerns has the resources either of manpower or finance to launch a really major attack on a new field. Among the smaller ones it means all too often, not only an absence of research and development effort within themselves, but also, in varying degrees, an inability to take advantage of the results of work of this kind being done in the Research Associations on their behalf. Following my participation in a visit to Russia in 1955, I included in an article published in the Institution *Journal* the question, 'How can we develop a more effective co-ordination of our own national resources without loss of the benefits associated with a free society?' This question is in process of being answered, but not with us as participants, in the collaborative arrangements now being made on the Continent as part of the European Free Trade Area negotiations. Agreements for joint research and development programmes, and for the co-ordination of other activities, are being worked out between groups of previously competing firms in different countries, and we may soon be faced with a very serious situation in the export field. These countries are no shorter of scientists and technologists than we are, if indeed they are as short. It therefore behoves the senior members of professions like our own to do all they can to ensure in their respective fields that what strength we have is not misapplied in unnecessary duplication, to the possible frustration of the young men who are joining us in our professional enterprises, and the detriment of our national wellbeing.

OUR RESPONSIBILITIES OVERSEAS, PARTICULARLY TO THE COMMONWEALTH

In the early part of my address I said that our scientists and technologists have become not only the guardians and guarantors of our national economy, but also the agents of a new diplomacy in which more than an expanding export trade is at stake. This something more is the improvement, undertaken for its own sake and not merely as an act of self-interest (though in fact this interest would be safeguarded automatically), of the distressingly poor economic and social conditions of vast portions of the world community.

In his Presidential Address entitled 'Technology and World Advancement' to the British Association for the Advancement of Science in 1957, Professor P. M. S. Blackett[8] said,

> In addition to maintaining its existing wealth, the Western world is saving and investing productively some 10 per cent of its average income of £300 per head (at 1949 prices); that is some £30 a year per head is being invested in additional plant and machinery to create more wealth. The pre-industrial countries of Asia only have about £20 a year per head to live on, that is for both consumption and production goods.

Professor Blackett's advocacy was for large-scale capital assistance to enable these underdeveloped countries to 'achieve the take-off towards sustained economic growth'. Mine will be for more educational assistance to the same end.

In fact the assistance which this country has already given within the Commonwealth since the war has been considerable—much more perhaps than is generally realized or for which we are always given due credit overseas. Thus the Colonial Office has provided through the Inter-University Council for Higher Education Overseas more than £12 million for university and university college development in Ghana, Hong Kong, Malaya, Malta, Nigeria, Rhodesia and Nyasaland, Uganda and the West Indies; and through the Council for Overseas Colleges of Arts, Science and Technology some £3 million for the establishment of these colleges in East Africa, Ghana, Nigeria, Sierra Leone, Singapore and the West Indies. In addition, the Commonwealth Relations Office has contributed to the Colombo Plan for assistance to South and South-East Asian countries close to £4 million for the provision of educational and research equipment, of expert advisers and of training grants. Members of The Institution will already be aware through an article published in the March, 1959, issue of the *Journal* of the latest offer of help under the auspices of the Plan, that of joint sponsorship by the Commonwealth Relations Office and the Federation of British Industries of a College of Engineering and Technology in the University of New Delhi.

In aggregate these contributions to the furtherance of education in the underdeveloped countries are second in amount only to those of the United States. On the other hand, Russia and West Germany have begun to play a more substantial part, and their sponsorship, for example, of the Institutes of Technology at Bombay and Madras, respectively, is on a particularly generous scale as regards both equipment and staff. Staffing is of crucial importance, and I am not a little anxious, for both absolute and comparative reasons, about this aspect of our undertaking in New Delhi.

However, the problem is a more general one, for it has proved, not only extremely difficult to staff from this country the institutions for which we have carried an initiating responsibility, but impossible to fill more than a fraction of the teaching posts overseas for which our help has been specifically sought. These difficulties have undoubtedly arisen from the shortage of teachers to meet the needs of our own educational developments. But I feel that they are also due in part to the lack of a clearly defined and unified plan for the secondment of teachers overseas as an act of national policy, carrying with it appropriate safeguards for their return, and an appeal to teachers—and not least to teachers of engineering—to devote a few years of their lives to this kind of international service. They would incidentally make a profound contribution to the future well-being of their own country.

There is a further way in which we have given great assistance, but in which we must also increase our contribution. We have good reason to be proud and thankful that so many of those now occupying senior positions in all walks of life within the Commonwealth and Colonies, and indeed in many other countries, came here as young men for their education and training. They have been coming during recent years in ever increasing numbers, under the auspices of the British Council and the Colombo Plan, of the Athlone Fellowship and the F.B.I. Overseas Scholarship Schemes, of their own Governments, and by individual initiative. This flow is now to be augmented by the large number of scholarships tenable in this country which the British Government offered at the Commonwealth Conference held in Oxford in July.

At the present time there are upwards of 40 000 overseas

students, including some 26 000 from the Commonwealth, in this country. Of the latter over 7 000 are university students, and of these about 1 600 are following undergraduate courses in technological subjects. No doubt a substantial proportion of the 2 600 post-graduate university students are also technologists. A further 6 500 from the Commonwealth are studying in the technical colleges.

In respect of their formal education we need, I think, have no anxieties about these young people. On the whole they make good and successful students and they exercise a valuable educational influence on our own. Moreover many, mainly voluntary, organizations are performing a supplementary, more informal but no less important, educational and social service on their behalf. But we in engineering are faced with the serious problem that it is becoming increasingly difficult to make arrangements in industry and the operating organizations for their practical training. This means that they are unable to obtain here what their own countries are least able to provide in the present early stages of industrial development and that on their return home they will be the less competent to contribute to it. Many of them are reluctantly taking advantage of opportunities being offered on the Continent, and they may well feel the more dissatisfied if the training which these opportunities afford is considered by the Institutions to be inadequate.

The difficulty in obtaining training facilities also applies to the more mature persons whom the Commonwealth Relations Office, the Colonial Office and the British Council wish to bring here to refurbish their professional skills and to learn the latest techniques.

I have already discussed the basic reason for this difficulty. Its solution is to be found, if it can be found, only in the acceptance by the smaller and more specialist firms of a greater measure of responsibility for providing good training facilities, either within themselves individually or by collaborative effort. In fact these firms have special merits for overseas students in that they compare more closely than do the large organizations with the type of manufacturing concern of which the industry of their countries is likely to be composed. It would be invaluable for these men to have the opportunity of training under proper guidance within a group of small firms, each with its characteristic production techniques, organizational structure and administrative methods, and of making comparisons between them in these respects.

Finally, there is one category of overseas visitor to which I attach special importance, that of members of staff of the engineering departments of the Commonwealth and Colonial universities and colleges. We must do all we can to help these men to establish high standards of professional competence and conduct in their own countries, and I hope therefore that industry will make a special effort on their behalf. I trust it will also be possible to accommodate increasing numbers of them in the technical teacher training colleges now undergoing expansion.

THE UNITY OF THE ENGINEERING PROFESSION

In most of the preceding I have referred to engineering in the general sense, because much that I have tried to say applies to each of its constituent branches. I now want to refer to two oppositely directed trends within its professional organization. One arises from an increasing recognition of the growing interdependence of these branches, and is directed towards closer collaboration between existing professional bodies and the avoidance, by this collaboration, of the creation of new ones. This has found expression in the joint committees of the Institutions of Civil, Mechanical and Electrical Engineers; in the establishment by these Institutions, The Institution of Chemical Engineers and the Institute of Physics of the British Nuclear Energy Conference and the subsequent incorporation of several other bodies as joint sponsors of it; and in the creation by some 40 institutes, institutions and societies of the British Conference on Automation and Computation. The recently announced proposal for amalgamation of the Institute of Physics and the Physical Society is a closely related step in the same direction.

The other trend arises from the desire of representatives of new fields of technical activity to secure distinctive recognition and status for them.

It is of great importance that the first of these trends should progress as rapidly as possible so that professional engineers may acquire a fuller sense of corporate responsibility in professional and public affairs, and achieve a collective status in the public mind more in accord with the importance of their national service. But its progress will be impeded if the initiative and vitality which characterize those associated with new developments are not given early, enthusiastic and satisfying recognition and support. Our own Institution is endeavouring to fulfil this responsibility towards its vigorous and rapidly expanding Electronics and Communications Section. The name of this Section was changed recently so as to lay emphasis on the growing importance of electronics within electrical engineering. This was expressive of The Institution's determination to afford full representation to those engaged in this branch of the electrical profession and to cater comprehensively for them. Consideration is now being given to further steps which may assist in these directions.

Anything which can be done towards unifying the profession at home while at the same time ensuring the full vitality of its constituent parts is of value in assisting our professional colleagues overseas. It is particularly important that our three major Institutions should hold analogous views on matters of education, training and professional recognition, since their joint representations to the Commonwealth Engineering Conference and at meetings of Eusec are of major significance to the progress of these international bodies.

THE ENGINEER—HIS DUE AND HIS DUTY IN LIFE

In my Introduction I referred to the responsibility which we senior members carry for ensuring that the newcomers to our profession are properly prepared by education, training, and example, for participation in its work. I did not intend the word *example* to apply only to our scientific, technological and administrative activities as professional engineers. My use of it was in fact a reflection of the inspiration which I derived from a paper read before the North-Western Centre of The Institution in 1925, the year of my graduation at Manchester University. This paper was entitled 'The Engineer—his Due and his Duty in Life', by Thomas Carter, Member,[9] and the recollection of its presentation and of the discussion which followed has remained with me over the intervening years. That evening I felt proud to be entering the electrical engineering profession. I have decided that I cannot conclude my address better than by quoting from the opening Section of this paper, which the late Mr. Carter headed 'The Rights and the Duties of Life' and which reads as follows:

> In these days, when many standards of the past are disappearing, when new thoughts are being conceived and new views are struggling for recognition, and when more than ever the safety of the future depends on the well-directed efforts of men of wide vision and clear understanding, it is not too much to say that unless each of us is contriving somehow to increase the common good, he were much better dead. . . .
>
> We are born into a family, into a nation, into the world; we

inherit from all the past; we are played upon by things around us, and we react on them in our turn; each one of us has his own qualities, his own defects, his own potential contribution to the service of his fellows . . . There are, then, two sides to a man's life, together making it complete: the one, his right to live; the other, his duty to the community . . . His right is to expect to be prepared in his youth for a job that he can do in his working days with enough comfort to leave his spirit unquenched if he grows to be old . . . On the other hand, the more we give every man a chance to reach his best, the more must he recognize that this good comes to him from the community, and the more must he strive while following his own bent, to avoid every activity that will put a brake on the wheels of progress. This, then, is a proper life: a constant giving and receiving, a perpetual interchange of services, ideas, and commodities, with an unfailing consideration of our neighbour's good as well as of our own, and a constant recollection that those who forget their common obligations, and merely demand their own due without caring what happens to others, are the slayers of the nation's soul. There is no right without its corresponding duty.

I hope this quotation may move many members of The Institution, the younger ones particularly, to read the paper and the ensuing discussion in full, and that it will help them, as it has helped me, to see more clearly their duties to the engineering profession and to the wider community of people, and of peoples, of which it forms a part.

BIBLIOGRAPHY

(1) Livingstone, Sir Richard: 'The Future in Education' (*British Association for the Advancement of Science Report of Annual Meeting*, 1936, p. 219.
(2) James, D. G.: 'Industry and the Universities', Conference of the Home Counties Master Printers' Alliance, October, 1956.
(3) Ashby, Sir Eric: 'Technological Humanism', *Journal I.E.E.*, 1958, **4**, p. 478.
(4) Ministry of Education Administrative Memorandum No. 570, 1957.
(5) 'The Supply and Training of Technical Teachers' (H.M. Stationery Office, 1957).
(6) Snow, Sir Charles [C. P. Snow]: 'The Two Cultures and the Scientific Revolution' (Cambridge University Press, 1959).
(7) Jackson, Sir Willis: 'The Partnership between Industry and Education', *Advancement of Science*, 1958, **15**, No. 58, p. 129.
(8) Blackett, P. M. S.: 'Technology and World Advancement', *ibid.*, 1957, **14**, No. 54, p. 3.
(9) Carter, T.: 'The Engineer: His Due and His Duty in Life', *Journal I.E.E.*, 1926, **64**, pp. 193 and 879.

CENTRE CHAIRMEN'S ADDRESSES

The Institution of Electrical Engineers
Abstract No. 3188
Jan. 1960

©

NORTHERN IRELAND CENTRE: CHAIRMAN'S ADDRESS

By T. S. WYLIE, Member.

'THE PROVISION OF POST OFFICE BROADBAND LINKS FOR COMMERCIAL TELEVISION AND OTHER SERVICES IN NORTHERN IRELAND'

(Abstract *of Address delivered at* Belfast 13*th October*, 1959.)

The extension of commercial television to Northern Ireland on the 31st October, 1959, has involved the Post Office in the provision of a microwave radio-relay system between Ballygomartin (Belfast) and Carlisle, a coaxial cable network in Belfast, and the installation of power, transmission and radio equipment at various stations along the route. The Ulster Television studio in Belfast and the I.T.A. transmitting station on Black Mountain are thus connected to the main I.T.A. network in Britain via the Post Office Network Switching Centre at Carlisle.

At the P.O. radio terminal, Ballygomartin, the received microwave signals working in the frequency range 3 800–4 200 Mc/s are converted to frequencies in the television video band (50 c/s–3 Mc/s) and transmitted along a 0·375 in diameter coaxial tube equipped for unbalanced video operation, via the Post Office Network Switching Centre, Belfast, and the Ulster Television studio to the I.T.A. transmitter on Black Mountain. The Black Mountain station will operate in the v.h.f. range on channel 9, band 3. The signals will be horizontally polarized to minimize interference with the London and Manchester stations, which use the same frequencies but vertical polarization.

The maximum capacity of the Northern Ireland microwave link using the available frequencies in the 4 000 Mc/s band is 12 r.f. channels, i.e. 6 in each direction. Five channels will be provided initially, one incoming to Belfast for television, one in each direction for broadband telephony, and a standby, or protection, channel in each direction. These standby channels are automatically switched into service when a fault occurs on a working channel, by means of baseband equipment installed at each radial terminal.

Mr. Wylie is at the Post Office Headquarters, Northern Ireland.

Microwave Link

One of the advantages of microwave propagation is that there is a greater total bandwidth available for the various r.f. channels. As television and multi-channel telephony groups require the transmission of a wide frequency band, microwave links are very suitable, and each r.f. channel is designed to take either 600 telephony circuits arranged in 10 supergroups working in the 60–2 540 kc/s frequency band, or, alternatively, it can be used for television having a bandwidth up to 5 Mc/s.

Although there appears to be no precise definition of 'microwaves', it is generally accepted that it means radio waves having wavelengths of 1 metre or less, i.e. the decimetre, centimetre and millimetre bands covered by the international classification. The Northern Ireland link equipment will use waves in the centimetre band of the order of 7·5 cm wavelength.

A characteristic of microwaves is that they behave like optical waves and travel along straight paths in a homogeneous medium; it is therefore necessary to site radio stations within optical distance of each other to work over a line-of-sight path. Belfast and Carlisle are outside optical range, and a study of ground contours between the two stations showed that three intermediate relay stations were required, situated at Enoch Hill, Cambret Hill and Riddingshill. The spacings range from 30 to 40 miles.

The optical laws of reflection and refraction can be used also to explain the behaviour of microwave beams; e.g. the average effect of the atmosphere is to refract the beams so that they tend to bend around the earth's curvature. Since the physical properties of the atmosphere (density, temperature, humidity) vary with height and with time, the amount of bending is not always constant.

Fading

Most radio links are subject to fading, the depth and frequency of which depend upon variations of atmospheric conditions which, in turn, depend to some extent upon the time of day and time of year; e.g. in winter there is less fading than in summer.

Microwaves are propagated by direct space rays, but reflections occur from the intervening ground and sea, and from layers in the atmosphere where relatively abrupt changes of refractive index occur. Direct and reflected signals arrive at the receiver aerial after travelling over paths of different length, and they are therefore of different phase relationship. Changes in meteorological conditions cause changes in the various path lengths, and the consequent phase changes produce variations in the amplitude of the resultant wave, thus causing a fade.

Fading is mostly of two kinds. One, which is the most prevalent and occurring most of the time, is due to interference variations between the direct and sea-reflected rays. The other kind of fading, which is worst on calm and sunny days, is caused by the formation of layers in the atmosphere which show a discontinuity in the vertical gradient of the refractive index of the atmosphere, so that additional reflected rays are produced. Turbulent air conditions associated with wind and rain can prevent the formation of these air layers.

The effects of fading are minimized by the use of automatic gain control in the radio receivers and also by the use of spaced diversity reception in the receiving aerial system. Spaced diversity is based on the observation that incoming signals of the same frequency received at different locations do not fade by the same amount at the same time. At Ballygomartin the main aerial is at a height of 45 ft and the diversity aerial is at 90 ft. The s.h.f. signals from these two aerials are combined and fed into a single receiver, a motor-driven phase shifter keeping the two signals in the proper phase relationship.

Aerials

An aerial on a point-to-point link serves two main functions. Firstly, it radiates the radio-frequency energy produced by the transmitter and fed to it through the waveguide. For maximum efficiency and power radiation, the impedance of the waveguide should match the impedance of free space. Secondly, it directs the radio-frequency energy along the wanted direction towards the distant receiver and minimizes radiation in other directions.

The type of aerial used in the Northern Ireland link is a paraboloid reflector, an advantage of this type being that it has a high degree of directivity for microwaves. The horn feed at the end of the waveguide is mounted diametrically across and at the focus of the paraboloid, so that a narrow pencil beam (parallel to the axis) is produced of width less than 2° between 3 dB points. The gain of such an aerial at 4 000 Mc/s is approximately 40 dB relative to an isotropic aerial, i.e. one radiating energy equally in all directions.

The reflector is made up from eight accurately machined cast-aluminium segments bolted together and to a central circular casting to form a paraboloid having an aperture plane of 10 ft diameter and a depth of 2 ft 6 in. A flat vertex matching plate fitted at the centre of the reflector serves to cancel out reflected power scattered back into the horn feed.

Waveguides

The idea of electromagnetic waves travelling through space is now commonly accepted. A further development is to think of these waves being passed along a hollow pipe or guide, their electric and magnetic fields being confined by the walls of the guide so that no radiation can pass out into space.

Waveguides are preferable to coaxial feeders for super-high frequencies as they have much lower attenuation. Generally speaking, coaxial feeders are used for wavelengths greater than 15 cm, as waveguides would be inconveniently large. For wavelengths less than 3 cm waveguides are essential, and between these two limits, coaxial cable may be used on short runs and where flexibility and ease of handling are required.

Precision mechanical engineering is essential in waveguide construction. Joints and bends must be perfectly aligned internally to maintain the wideband properties of the guide, and also to avoid irregularities which might produce reflections and generate spurious modes of propagation. Coupling between sections is effected by means of flanges brazed to the ends of each section. Alignment is by means of dowels fitted in the face of the flange, and a gastight, weatherproof joint is ensured by fitting Neoprene gaskets between the flanges. It is very important to prevent moisture getting into waveguides, since any small particles of liquid can cause losses due to scatter or absorption, and condensation of water vapour will produce an attenuation exceeding 10 dB. Protection is obtained by circulating dry air through the external waveguide runs, obtained as a separate feed from the blower unit which has been provided for the forced-air cooling of the microwave valves in the radio terminal cubicles. Silica gel is used in the drying unit. Within the building, protection is obtained by means of an internal coating of gold plating.

Coaxial Cables

A coaxial pair consists of a copper centre conductor, 0·104 in diameter, and an outer conductor of soft copper tape formed into a tube of 0·375 in internal diameter. The inner conductor is held centrally within this tube by means of polythene discs fitted over the inner conductor at 1·3 in intervals. Two mild-steel tapes are applied in succession over the copper tube to give it mechanical strength, while interstitial pairs used for audio-frequency control circuits, etc., maintain the circular cross-section of the cable. The lead sheath is protected from corrosion by an external lapping of hessian tape, and sections on the mountain slopes are laid directly in the ground and protected by helically wound steel wires. Anti-creepage devices have been fitted to the cables laid along the steeper slopes.

The cable system has been pressurized to 9 lb/in^2 with compressed dry air. This not only gives a warning of sheath faults, but serves to maintain insulation during fault localizing. The network is divided into three independent gas sections, having contactor gauges which operate when the pressure falls to 7·5 lb/in^2 and complete an alarm circuit.

Sound Programme

The sound programme is not transmitted over the radio link with the video signals but is routed on a Belfast–Port Patrick carrier group. For good-quality reproduction, the audio programme channel as recommended by the C.C.I.T.T. should have a frequency range extending to 10 000 c/s. A carrier-derived channel has been utilized by installing special equipment at Belfast and Port Patrick which gives a bandwidth of 30–10 000 c/s derived from the bandwidth normally used for three audio channels. The 'music' band is frequency translated by means of a metal-rectifier suppressed-carrier modulator fed

with a 96 kc/s carrier. The lower sideband, 86·6–95·97 kc/s, is added to the sideband frequencies from the remaining nine channels of the basic carrier group working in the range 60–108 kc/s.

Power Supplies

A continuity set and standby engine have been installed by the Post Office, so that a 'no-break' a.c. supply kept within strict limits can be fed to the various racks of equipment. The continuity set consists of a d.c. motor, 3-phase induction motor and single-phase alternator, coupled together and bolted to a common bedplate. The standby set consists of a 3-phase alternator driven by a Diesel engine. The mains supply normally drives the induction motor coupled to the single-phase alternator supplying the station load. Under normal conditions, the d.c. motor is running idle with only the field energized. When the supply voltage falls below a certain value, or is interrupted, a quick-acting contactor changes over the drive from the a.c. motor to the d.c. motor operated from station batteries. At the same time the Diesel engine starts up, and when it has run up to speed the 3-phase alternator coupled to it provides the supply to the 3-phase induction motor, and the d.c. motor is cut out. When the mains supply is restored the drive is automatically switched back to the a.c. motor. A full-wave selenium rectifier keeps the motor battery fully charged.

Equipment

Frequency modulation of the r.f. carrier is a feature of the system, and this method has important advantages over amplitude modulation, e.g. improved signal/noise ratio, reduced distortion due to amplitude non-linearity, etc. The equipment provides for unattended operation, and at the Belfast and Carlisle radio terminals, facilities are provided for remote control, and visual displays show the state of the equipment and power supplies at the intermediate stations.

A reflex-klystron oscillator is used in the transmitter to provide the s.h.f. radio signal. The video-frequency band is applied to the reflector cathode circuit of the klystron to give a maximum frequency deviation of the s.h.f. carrier of 8 Mc/s peak-to-peak for television.

The frequency-modulated output signal is translated to an intermediate frequency of 70 Mc/s, and after amplification, to the radiated s.h.f. carrier frequency. Amplification at the final super-high frequency is obtained with a travelling-wave tube which combines extremely broadband characteristics with high gain, and the final power output is 5 watts. Automatic frequency control of the klystron is achieved by comparing its frequency during the synchronizing-pulse period with that of a crystal-controlled source and arranging for any error to be corrected by mechanical tuning of the klystron cavity via a small motor. At the intermediate repeater stations and at the terminal receiver, the incoming microwave signal at a level of approximately 1 mW is applied to a crystal diode mixer, together with an s.h.f. signal derived from a temperature-controlled coaxial-line oscillator to produce a 70 Mc/s intermediate-frequency signal. At this frequency most of the amplification is provided to compensate for the overall path attenuation. An automatic-gain-control amplifier controlled by the level of the incoming signal compensates for fading up to approximately 30 dB.

At repeater stations, the i.f. signal is changed back to the s.h.f. band and amplified in a travelling-wave tube for onward transmission.

At the receiver terminal the i.f. signal is applied to amplitude limiters and to the frequency discriminator. This comprises frequency-selective circuits, the outputs of which are detected by crystals and fed to the negative-feedback video amplifier.

Conclusion and Acknowledgments

The Northern Ireland radio link will be the first Post Office internal system in the United Kingdom to provide facilities for the transmission of both telephone circuits and television. The part played by the Post Office in providing this link illustrates the varied range of engineering techniques which have been necessary.

The Institution of Electrical Engineers
Abstract No. 3183
Jan. 1960
©

MERSEY AND NORTH WALES CENTRE: CHAIRMAN'S ADDRESS

By T. A. P. COLLEDGE, B.Sc.(Eng.), Member.

'THE DEVELOPMENT OF THE TELEPHONE SYSTEM IN GREAT BRITAIN'

(ABSTRACT *of address delivered at* LIVERPOOL *5th October*, 1959.)

The telephone when first invented by Bell consisted of a steel clock-spring attached to a piece of goldbeater's skin. The movement of the spring in front of a magnet caused a varying magnetic field and in consequence a varying electric current in coils around the magnet. Both transmitter and receiver were of the same construction.

Until very recently the receiver still employed the same principle as Bell's, but with many improvements in detail. To-day a new type is being introduced known as the 'rocking-armature' receiver. This consists of a shallow conical diaphragm mechanically coupled to one end of a small bar magnet lying in the plane at right angles to the axis of the cone. The magnet is pivoted at the centre and rocks about this pivot under the influence of current in coils positioned near the poles of the magnet. The old-type receiver suffered from the compromise that had to be made in order to get a satisfactory balance between acoustic, mechanical and magnetic properties of the diaphragm. In the new type the conical diaphragm has better acoustic properties and can be made of non-magnetic materials having optimum mechanical properties. In addition, a magnetic circuit of much greater efficiency can be designed independently of the diaphragm.

In the case of the transmitter, Professor Hughes invented a microphone employing the property that varying pressure between two carbon blocks caused electrical resistance, and in consequence the current in the circuit, to vary. This property of carbon was used by Hunnings, who invented the carbon-granule transmitter in 1879. This contained a multitude of loosely packed hard-carbon granules in series in an electric circuit. The pressure on the granules was made to vary by a metal diaphragm on which sound waves impinged. The construction of the modern transmitter has undergone considerable change in detail, but the principle remains the same.

Mr. Colledge is in the Post Office Engineering Department.

In 1876 the first telephone was put into commercial use. A problem immediately arose, since it would be impracticable to run a pair of wires from each customer to every other customer. This was resolved by the invention of a switchboard, the first of which was put into operation in 1878 in the United States. The first exchange in Great Britain was opened in London the following year.

As the size of the exchange grew the number of customers increased beyond the capacity of handling by a single operator. This problem was solved by Mr. Hawes of the United Telephone Company, who devised a multiple switchboard, on which each customer's connection is repeated at regular intervals, the repetitions being such that any operator can reach all the customers' connections. The first vertical switchboard to be erected in this country was in the Dome of Liverpool Cotton Exchange in 1884.

Apart from their extension in size these manual exchanges have undergone very little change in the method of operation. The early types required magnetos for calling. Various intermediate steps led to the final advance when a central battery was installed at Bristol in 1900. When a customer lifted his telephone he automatically gave an indication to the operator that he was calling. This perhaps can be regarded as the first step towards automation.

The first automatic exchange was opened at Epsom in 1912; it worked on the Strowger system. In the next decade or so a number of other types of automatic exchange were installed throughout the country: a Lorimer exchange at Hereford, 1914; a Western Electric rotary system at Darlington, 1914; a Siemens step-by-step at Grimsby, 1918; and a Relay automatic system at Fleetwood, 1922. This decade may be regarded as the trial period; the outcome was the adoption of the Strowger system as a standard for this country. The decision to standardize had its repercussion on industry, on training and on organization; it clarified the position for many years.

In its main essentials the system is simple, is easy to understand and to observe, and still has attraction on this score. Every customer is connected to a simple signalling unit which connects his line to the common connecting equipment, or selectors. This can be achieved in two ways. Either each customer is connected directly to a switch (a uniselector) which finds a free outlet to the next group of switches, or the customers are connected to banks on groups of switches (line-finders). In the latter case, on receiving the caller's signal, a free line-finder, which has been preselected by an allotter, searches for the customer's line and extends it to the next group of switches.

The method to be used is decided on an economic basis. Where the number of telephone calls is high the uniselector type of exchange is the best, and where the calling rate is low the line-finder type of exchange is more economical. Modern exchanges are designed to have a mixture of the two systems.

In large towns a problem arose with regard to the connections between exchanges, a very similar one to the initial problem of the connections between customers. This was resolved by the introduction of the director system with a tandem exchange at Holborn, London, in 1927. Each exchange in London was given a code of three letters which preceded the customer's number. This code was the first three letters of the exchange name.

The director in any exchange will accept the code dialled and translate it into as many as six trains of impulses. It will send out these trains of impulses whilst receiving and storing the required customer's number. The trains of impulses will route the call via other exchanges if necessary and set up the circuit to the required exchange. The director will then pulse out the required number and on completion will be released from the circuit to translate and set up other connections. The total time it is in use for setting up a call is about 18 seconds. The director system enabled standard switches and apparatus to be used, reduced the number of digits to be dialled and, what was more important, standardized the code to be dialled for any exchange throughout the director area.

In contrast to these large automatic exchanges are the small unit automatic exchanges, called U.A.X.'s, which serve small and remote districts. The first of these was installed in 1929 and was designed to provide a 24-hour service. They have been developed to provide inter-dialling facilities to and via their parent exchanges. They are unattended and, where electric power is available, work on the battery-float system. Otherwise the batteries have to be charged by a prime mover, generally a petrol engine and dynamo.

The maintenance of an automatic exchange, with its complicated circuits and its numerous relays and switches, itself presents a problem. The automation of this function, too, is logical and essential if costs are to be lowered and a good grade of service is to be maintained. Apart from fault clearing, such maintenance should be of a preventive nature to locate and rectify faulty apparatus before it seriously affects the service. To achieve this, periodic application of tests to switching equipment are necessary, the periodicity being varied according to the importance of the equipment. On the introduction of automatic telephony this functional testing was done for many years by a maintenance engineer operating a portable tester which was connected in turn to each item of apparatus to be tested—a method which is still used in small exchanges.

In the late 1920's, automatic routiners were brought into use. They were automatically connected and they automatically applied the tests. When any item failed to pass the test they indicated the faulty item and gave an alarm. The maintenance engineer set the routiner on and dealt with the fault. Later types of routiner automatically record details of the fault.

Since the war, a through-call routiner, or artificial traffic equipment, has been developed and installed at the larger exchanges. Transportable models have also been made for smaller exchanges. This equipment passes calls automatically between 24 spare calling equipments and 25 spare numbers on an exchange in a complete cycle of 600 different combinations. When a call fails an alarm is given for maintenance attention. This machine has the great advantage that it tests all the items of equipment that customers meet in setting up their calls. It is not so critical as the automatic routiners, which generally apply limit tests.

By 1939 there were 2768 manual exchanges and 2925 automatic exchanges catering for over two million exchange connections. Plans were already in hand and studies were being made for the full automation of the telephone system.

Automation involved, not only the need for each customer to be able to dial any other customer, but the need to record the correct charge. Normally charges were made from exchange to exchange, and automatic metering of calls up to four units (15 miles) was introduced in 1935. Calls beyond 15 miles had to pass via an operator, who then recorded them, and they were eventually priced and debited to the correct account.

Group charging was introduced in January, 1958. This meant that charges were made from about 600 fixed points instead of 6000. The area of unit fee was considerably extended in all cases and cut down the need for the interposition of an operator for all calls to surrounding areas. It was a further step towards automation. The need for multi-metering on local calls was virtually eliminated, and operators were reduced in numbers. It was an essential preliminary to the introduction of subscriber trunk dialling.

In subscriber trunk dialling a system of national telephone

numbers will be used. It will be based on each customer's present number, and a robot operator will interpret the numbers dialled by a customer and route the call. The charging of such calls will also be done automatically and recorded on the customer's meter. Such charges will be based on time and distance.

Experiments are now going forward with electronic exchanges, in which the electro-mechanical switches are replaced by electronic switches. The advent of the transistor may eventually make all mechanical switching at automatic exchanges obsolete. There is no theoretical reason why any two customers cannot be connected together by electronically controlled switches of the thermionic-valve type. This is being tried at present but the power consumption is very high. The very low power consumption of the transistor and its extremely small dimensions makes it a very attractive alternative. We appear to be entering another phase of telephone development akin to the decade 1912–22, when the various automatic telephone systems were being tried out.

One further word with regard to human operators. On the final completion of subscriber trunk dialling, we shall still need to be able to call an operator for emergencies (police, fire, ambulance), for such services as personal calls, reverse-charge calls, etc., and provision will be made for this.

So far, no mention has been made of the connections between exchanges or of the connection from the customers to their exchanges.

The first customers were connected to their exchanges by overhead wires. As the number of customers grew the wires had to be supported on roof standards, which became enormous structures supporting both wires and cables. Underground cable soon became essential. Generally now ducts are laid and cables drawn through them. This means that plans must be made for a period of 20 years so that public highways do not have to be disturbed too often. Such plans are based on forecast figures for the penetration of the telephone into the areas concerned and will, of course, vary with the type of property.

The cables that are drawn into the ducts will be planned for shorter periods, since they can easily be augmented and readily replaced if damaged by corrosion or other hazards. When one realizes that some of the customers' cables in the big towns may contain as many as 1 800 pairs of wires, it is clear that an expeditious renewal is very necessary in case of faults. The Post Office, like other undertakers, suffers from cable-sheath corrosion. In the past the majority of cables have been sheathed in lead, but experiments are going ahead with cables sheathed in other materials the most promising of which appears to be the plastic sheathing.

In order to make full use of the local cables, cross-connection points (cabinets and pillars) are provided at junctions where the 20-year-sized cable to customers' distribution points meet the main cables back to the exchange. This provides flexibility, and full use can be made of spare pairs until the main cables are in full use, when they can be augmented.

A further development is now in the field-trial stage in the form of a line connector, where a small portion of the exchange is put into a container over a mile from the exchange. Four rotary line switches connected to four cable pairs to the exchange can cater for up to 22 customers with a saving of 18 pairs to the exchange. It is a type of decentralization, but the economics of this development are being watched as they are of a border-line nature.

The local cables to exchanges have changed considerably in size and in gauge of conductor. The maximum size of cables has gone up and the size of the conductors has decreased, the smallest now being 4 lb per mile of wire. This has resulted in an economy in duct space and in the saving of copper in the conductors. It has also meant an alteration in the terminal equipment in the exchange. With modern equipment the loop resistance of the line between the exchange and the customer's telephone can now be as high as 1 000 ohms, whereas with the older manual exchanges such lines could not have a greater resistance than 350 ohms.

Owing to the development in transmission technique, the provision of circuits between exchanges has undergone radical change and has resulted in considerable economies. The Engineer-in-Chief of the Post Office recently quoted that in 1915 three circuits erected overhead between New York and San Francisco required 2 700 tons of copper. The London–Birmingham underground cable of 52 pairs used 760 tons of copper, and even more copper would have been used but for the use of loading coils, introduced in 1901. With present-day types of construction we would expect to obtain about 1 000 telephone circuits for this amount of copper.

The manufacture of telephone cables calls for a great deal of precision so that the electrical characteristics of all pairs of wires are as nearly the same as possible, in order that the current in one pair does not cause induced current in any other pair.

Generally the wires in underground cables are in groups of four, called a quad, so that each quad contains two pairs. The quads in turn lie round the centre, and additional layers are added to give larger cables. The wire conductors are copper and are insulated with paper wound around each wire in a helix. These are generally known as audio cables.

The advent of the thermionic valve made possible the invention and introduction of the telephone repeater in 1920. This enabled light-gauge conductors to be used for audio circuits. A further advance was the development of carrier systems and the saving of repeaters by mass amplification of groups of channels by a single repeater. Finally the coaxial tube was evolved to take over a thousand circuits. This type of cable consists of a conductor held centrally in position (by non-conducting discs spaced throughout the length of the cable) relative to an outer copper helical tape, the whole being insulated and contained in a lead- or plastic-sheathed cable. Intermediate repeaters amplify the whole range of frequencies and are spaced at 6 miles. Shorter spacing will increase the number of circuits in such a cable.

In audio cables each pair of wires is designed to transmit current in the speech band having a frequency range of 300–3 400 c/s. In carrier working each speech band is transposed to a different position in the frequency spectrum. Normally 24 channels are arranged in the frequency range 12–108 kc/s. Such a band is known as a 'group' and is transmitted over a single pair of wires, whilst on a coaxial cable complete groups are further transposed in the frequency spectrum of 60–2 540 kc/s. Such a spectrum will accommodate 600 channels. Carrier and coaxial cables form the backbone of the trunk network. Audio cables are, in general, used only for customers' lines and the comparatively short junction circuits between neighbouring exchanges.

In a further development, which is still in the experimental stage, transistor repeaters are being used; these are smaller and require much less current.

It will be seen that, in the development of the telephone system, use is made of technical developments as they arise. This has been made possible by the continued close co-operation between the Post Office Research Department and the manufacturers of telecommunication apparatus and cables. It ensures that all possible avenues for progress are being explored. It results in advantage being taken, at an early date, of new inventions and techniques to improve our telephone facilities on a practical and economic basis.

The Institution of Electrical Engineers
Abstract No. 3189
Jan. 1960
©

SOUTHERN CENTRE: CHAIRMAN'S ADDRESS

By W. D. MALLINSON, B.Sc.(Eng.), Member.

'WEAPON SYSTEMS ENGINEERING'

(ABSTRACT *of Address delivered at* PORTSMOUTH, *7th October*, 1959.)

Systems engineering is the process of optimizing each element of the system in relation to the others for its function in accomplishing the task required of the system. The key to success is in the qualities of the systems engineer, who must be a combination of mathematician, physicist, engineer, artist and advocate. The steps he has to take in engineering a weapon system are the feasibility study, the appointment and knitting together of design authorities, the preparation of design specifications and programmes, the statement of limitations, and the evaluation plan. The salient features of these are discussed, with special emphasis on the need to complete a project quickly instead of seeking perfection at the cost of delays. Steps towards achieving this are to telescope the engineering design into the experimental stage, to analyse the results of equipment and system trials simultaneously with taking the records, and to telescope production into the development stage.

[The Address was illustrated with pictures of equipment forming part of weapon systems, and closed with a film of a weapon system undergoing trials.]

Mr. Mallinson is at the Admiralty Surface Weapons Establishment.

DISCUSSION ON 'THE RELATION BETWEEN PICTURE SIZE, VIEWING DISTANCE AND PICTURE QUALITY'*

NORTH-WESTERN ELECTRONICS AND COMMUNICATIONS GROUP, AT MANCHESTER, 12TH NOVEMBER, 1958

Mr. S. D. Mellor: The author has added to the difficulty of having his conclusions accepted by introducing the somewhat artificial conception of viewing distance as a measure of picture quality. It would have been more convincing if the conclusions had been supported by the results of a test requiring the observers to make direct comparisons, e.g. between (*a*) a 405-line 3 Mc/s picture and (*b*) a 405-line 1·5 Mc/s picture, or between (*a*) a 625-line 3 Mc/s picture and (*b*) a 450-line 3 Mc/s picture.

The inference from Fig. 6 is that the majority of observers would have been unable to detect any difference in picture quality on the first test, but that, on the second test, the improvement would have been outstanding. If these results had been obtained and if the sample size had been adequate, the author's critics would have been more effectively silenced. Possibly such tests have been carried out. If so, would the author disclose the results?

Mr. L. C. Jesty (*in reply*): Mr. Mellor's criticism of the paper is admitted in Section 5 of the paper. Unfortunately it was not possible to carry out the additional experiments he suggests at the time. It will become immediately clear, however, to anyone who cares to try it that viewing distance and apparent picture sharpness are closely related. An 'unsharp' picture becomes 'sharp' very suddenly as viewing distance increases (as indeed our observers recorded). The point at issue, of course, is whether two different kinds of picture which give the same chosen viewing distance will also appear to be equally acceptable (sharp) when viewed from this chosen distance. I think all the circumstantial evidence available shows that they would, but as Mr. Mellor points out, it would be better to establish this experimentally.

However, the results are not useless without this further evidence. It is quite clear, for example, from Fig. 6 that, in a given size of living room suitable for a 21 in receiver reproducing a 405-line 3 Mc/s picture, a 27 in receiver displaying a 625-line 3 Mc/s picture could be substituted, since they both have the same 'chosen viewing distance'. (I do not really think there is much doubt which picture would be preferred.)

* JESTY, L. C.: Paper No. 2540 R, February, 1958 (see 105 B, p. 425).

The results of a related experiment may be of interest and are given in Fig. D. In the recent 625-line u.h.f. broadcasting tests (summer, 1958) a number of dual transmissions of the same programme material were made on the 625-line experimental channel and on the regular 405-line Channel 1 transmitter. We invited teams of observers to view these on two 27 in monitors. They were asked to grade the two pictures (marked A and B) with respect to each other, by voting for one of the seven grades shown in Fig. D. Apart from one particular session (1st August, 1958) when an obvious fault (which could not be located) was present on the 625-line picture, the preferred choice was obviously between 'better than' and 'much better than'.

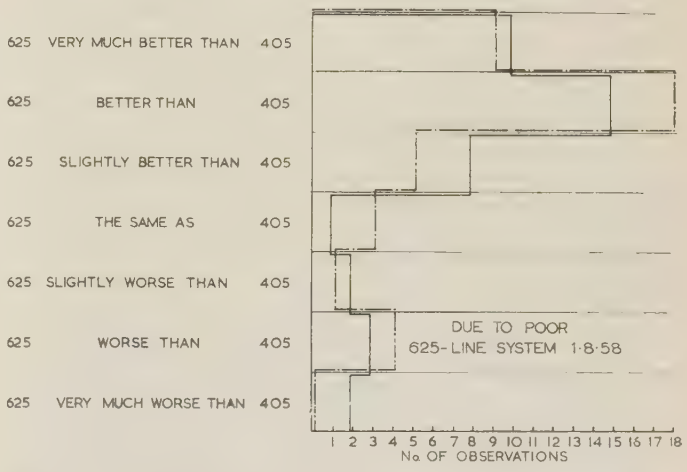


Fig. D.—Observer grading of 625- and 405-line television reception viewed on 27 in monitors 1st–12th August, 1958.

—·— 30 television engineers. 40 observations.
—— 27 other observers. 41 observations.
10 transmissions: 9 film, 1 still picture.

DISCUSSION ON 'A TRAIN PERFORMANCE COMPUTER'*

NORTH-WESTERN MEASUREMENT AND CONTROL GROUP, AT MANCHESTER, 17TH FEBRUARY, 1959

Mr. A. M. Lyall: There are in existence three models of the computer described in the paper. My experience has been on the prototype machine and the first production model, which employ the type of transducer incorporating uniselectors. These are so noisy that it proved impossible to use the machine in a general office during working hours. A separate office is therefore required, which can have the additional advantage of being made dustproof.

A further disadvantage of the uniselector transducer is the fact that it is non-directional and cannot sense the direction of rotation of the integrating meter. A phase-sensitive circuit is therefore required which must be capable of operating at both low and high currents. This gave some trouble in the initial stages, but it is now satisfactory.

The noise of the uniselector, however, gives the operator the impression of actually driving the train, since the frequency of the 'clicks' increases as the speed rises, and decreases as it falls. A large proportion of the time taken to perform a given run is taken up by the preparation of the gradient tape. Some of this time could be saved by the use of an electrically driven punching machine with a mileage indicator which could be stopped at each gradient change. The punching could then be carried out in one operation instead of each hole having to be punched separately as in the present hand-driven punching machine.

The paper mentions an error of $\pm 1 \cdot 5\%$. This may seem rather large, since it represents $6\frac{1}{2}$ min on a run from London to Glasgow, but it must be remembered that it is not possible to estimate train resistance to such a degree of accuracy. A cross wind, for example, may so increase the friction between the wheel flange and rail that the train resistance is raised by much more than $1 \cdot 5\%$. In practice, this is catered for by the 'make-up' time in the train schedule.

The paper mentions one method of allowing for speed restrictions. An alternative method, especially in the case of multiple-unit stock where the driver has a very limited number of notches on his controller, is to apply power until the restricted speed is reached, and then coast until the speed falls a little, apply power again to restricted speed, and so on, as the driver of the actual train would, in fact, do.

Mr. G. R. Higgs: A complicated computer may seem unnecessary, but it should be realized that some railways require a very large number of train-performance curves to be submitted with a tender. In some cases as many as 100 man-days can be consumed in making these run curves, and this, of course, applies to each of the several tenderers involved. In these cases, particularly, the machine is very useful to a contractor.

I should be interested to know what plans, if any, the authors have for making the device even more automatic. It would be very convenient if, having once been set up to compute a 4-hour train journey, it could be left severely alone for the four hours, coping automatically not only with the full-power running, which it already does, but also with the reduced-power running involved in the observation of speed restrictions.

It has been demonstrated that a saving can be effected by using the same gradient chart for both directions on the run. Presumably compensated grades have to be 'decompensated' in order that the curve allowance can be used in the opposite sense for the other direction of travel. This might nullify the saving.

Taking into consideration the accuracy of the machine and the data fed to it, it is worth noting that a 12-coach train at 60 m.p.h. experiences about half of its total train resistance in the form of air resistance. A 20 m.p.h. head wind would nearly double the air-resistance portion, and increase the total resistance by something over 20%. This shows the futility of striving for extreme accuracy in the computer itself.

Prof. E. Bradshaw, and **Messrs. M. Wagstaff** and **F. Cooke** (*in reply*): Mr. Lyall rightly comments on the advantages of the later type of disc-rotation transducer shown in Fig. 5(*b*) of the paper. He also draws attention to the refinements which could be introduced into the gradient type of punch. We can see no reason why his suggestions could not be adopted.

We can confirm that the suggested method for dealing with speed restrictions is not only practicable but also convenient.

Messrs. Higgs and Lyall both put the accuracy of the computer and of the available data into the proper perspective. We appreciate their comments on this matter, but we would nevertheless like to repeat what we stated in our reply to the London discussion, namely that the estimate of the error limits of $1\frac{1}{2}\%$ was very conservative and that a much better performance is obtained in practice.

Whilst one is tempted continuously to refine the device, we have no plans to achieve the fully automatic state suggested by Mr. Higgs. It is of interest, however, that each user of the machine has made detailed refinements to suit his own needs and inclinations. The form of the computer, depending so much on low-power electrical circuits, lends itself to such modifications as suit individual needs.

Because the use of the same gradient chart in both directions involves punching the chart at the beginning and end of a grade, there is no difference in the actual number of holes to be punched compared with the use of a separate chart for each direction. Compensated gradients involve some additional work in the coding of the chart from the information given on the track profile, the amount being the same whether separate charts are used for each direction or not. The saving in time resulting from using one gradient chart for both directions is in the marking out of the chart and the elimination of one passage through the punch, which represents an appreciable portion of the total preparation time.

* BRADSHAW, E., WAGSTAFF, M., and COOKE, F.: Paper No. 2425 M, November, 1957 (see 105 B, p. 560).

621.396.67: 621.396.8.029.

The Institution of Electrical Engineers
Paper No. 3079 E
Jan. 1960
©

THE PERFORMANCE OF A BALANCED AERIAL WHEN CONNECTED DIRECTLY TO A COAXIAL CABLE

By G. D. MONTEATH, B.Sc., A.Inst.P., and P. KNIGHT, B.A., Associate Members.

(*The paper was first received* 3*rd March, and in revised form* 12*th June*, 1959.)

SUMMARY

Measurements have been made of the radiation patterns of typical aerials, such as would be used for domestic reception in the v.h.f. band, connected directly to a coaxial cable without the intervention of a balance/unbalance transformer. The vertical radiation pattern may be appreciably distorted, owing to the excitation of currents on the outside of the cable and the sensitivity to vertically-polarized interference originating near to and below the aerial may be increased. The distortion of the radiation pattern appears to be less serious with directional aerials. The use of a balance/unbalance transformer or a balanced feeder would be desirable, but unless the input circuit of the receiver is either balanced, or screened and unbalanced, no ideal solution exists.

(1) INTRODUCTION

The paper is concerned principally with aerials used for domestic reception in bands I (41–68 Mc/s), II (88–95 Mc/s) and III (176–216 Mc/s). The experimental work has been done with a vertically-polarized aerial, but some of the conclusions are applicable to horizontally-polarized aerials used for receiving television and frequency-modulated sound transmissions.

V.H.F. receiving aerials sold to the public are usually balanced, and no balance/unbalance transformer, or 'balun', is provided. Nevertheless, the feeder is usually coaxial—almost invariably for television and as often as not for f.m. sound. It is commonly believed that a coaxial feeder is less sensitive to local interference than an unscreened twin feeder, but this may not be true.

The object of the investigation described was to determine whether the use of coaxial feeder can impair the performance of a balanced aerial appreciably.

(2) THEORETICAL CONSIDERATIONS

(2.1) General

Fig. 1 shows balanced dipoles, oriented for the reception of vertically and horizontally polarized transmissions, connected to both coaxial and balanced feeders. Apart from its intended function—to convey signals from the dipole to the receiver—the feeder can influence the performance in two ways. First, currents induced on the feeder will result in reradiation; in other words the feeder acts as a parasitic* element. This will happen for any type of feeder, but may be avoided when the dipole is horizontal by disposing the feeder in the equatorial plane of the dipole.

A coaxial feeder [Figs. 1(*a*) and (*b*)] will also act as a driven* element, currents excited directly on the outside of the feeder passing partly into the inside at the connection to the dipole. This will occur whether the dipole is horizontal or vertical, and for each case both vertically- and horizontally-polarized signals

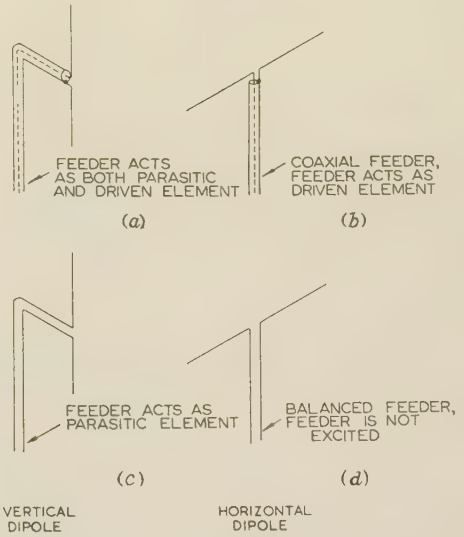


Fig. 1.—Dipoles with balanced and unbalanced feeders.

can be received as a result of the excitation of currents on the vertical and horizontal parts of the feeder respectively.

A coaxial feeder connected to a vertical dipole will act as a parasitic element and a driven element simultaneously.

(2.2) The Feeder as a Parasitic Element

An estimate of the order of magnitude of this effect can be obtained by assuming as a first approximation that the current excited on the vertical portion of the feeder will be much the same as if the feeder were replaced by an infinite vertical conductor. This assumption ignores reflections at the ends of the feeder. Carter's method* is used to calculate the radiation pattern of a vertical dipole parallel to an infinite vertical rod and the contribution due to reradiation from the rod is halved. This approach leads to the conclusion that the effect of the feeder as a parasitic element is not very great; such effect as it has is that of a reflector. The current induced on the feeder varies in a logarithmic manner with its diameter, and is not therefore affected greatly by the type of feeder used. Nevertheless, a thin feeder has rather less effect than one which has been effectively thickened by bonding to a metal supporting pole. The effect of a pole not bonded to the feeder cannot be predicted easily, since the pole is usually too short for resonance to be neglected. This case will not be considered further.

(2.3) The Feeder as a Driven Element

It is convenient to treat the aerial as for transmission and to apply the principle of reciprocity when its receiving properties are considered. The transmitting aerial and feeder in Fig. 2(*a*)

* Strictly speaking, the use of the words 'parasitic' and 'driven' implies transmission rather than reception, but the reciprocity principle ensures that the radiation pattern is the same when receiving as when transmitting.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Messrs. Monteath and Knight are with the British Broadcasting Corporation.

* CARTER, P. S.: 'Antenna Arrays around Cylinders', *Proceedings of the Institute of Radio Engineers*, 1943, **31**, p. 671.

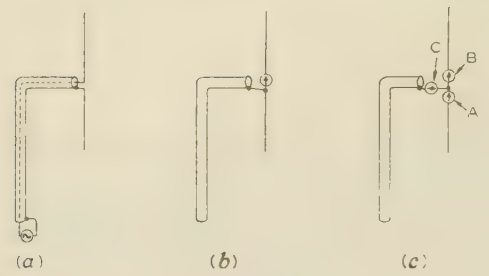


Fig. 2.—Equivalent circuit.

(*a*) Transmitting dipole fed by coaxial cable.
(*b*) Transmitter replaced by elevated generator of zero internal impedance and e.m.f. $2V$.
(*c*) Generator replaced by three equal generators of e.m.f. V.

have the same radiation characteristics as the equivalent circuit shown in Fig. 2(*b*). Applying the superposition theorem, the generator shown in Fig. 2(*b*) can be replaced by three generators of equal e.m.f., as in Fig. 2(*c*). Two of these, A and B, excite the dipole in a balanced manner, but the principal effect of generator C is to excite currents on the outer conductor of the feeder. It will also excite currents in opposite directions on the limbs of the dipole, but the resulting radiation will be small, the dipole acting as a half-wave 'earth bar'.

The current excited on the outside of the cable by generator C takes the form of travelling waves directed downwards towards the transmitter. There will also be upwardly-directed waves resulting from reflection at the bottom of the feeder, but these will be smaller in amplitude. Current flowing on the horizontal part of the feeder will result in horizontally-polarized radiation. The waves travelling downwards on the vertical part of the feeder will result in vertically-polarized radiation mainly directed obliquely downwards. Radiation directed obliquely upwards will be caused both by the reflection of downwardly-directed radiation at the ground and by direct radiation from the waves of current travelling upwards on the feeder.

Generators A and B in Fig. 2(*c*) will together be loaded by the impedance of the dipole, which is normally adjusted to about 70 ohms. The load presented to generator C will approximately correspond to the impedance of a single-wire travelling-wave aerial. It is probably mainly resistive and of about 250 ohms. It follows that the power radiated as a result of the excitation of the feeder as a driven element will be only some 7% of the total power. Nevertheless, this could have an appreciable effect on the radiation characteristics of the aerial. The effect will be much greater if the aerial is mismatched so as to have a low conductance, or if the feeder is short and resonant.

(2.4) The Resultant Effect of the Feeder

Expressing the conclusions reached in Section 2.3 in terms of a receiving aerial, and combining the effects of the feeder as a driven and as a parasitic element, the total effect to be expected is as follows:

Balanced Feeder, Horizontally-Polarized Aerial.—The feeder will have no effect.

Balanced Feeder, Vertically-Polarized Aerial.—The feeder will act to some extent as a parasitic reflector, but the effect will not be large.

Coaxial Feeder, Horizontally-Polarized Aerial.—The feeder will act only as a driven element. Reception on the horizontal portion, if any, will distort the horizontal radiation pattern. This may become asymmetric and the zeros may be filled in. Owing to reception on the vertical part of the feeder, the aerial will be sensitive to vertically-polarized waves, particularly if these are radiated obliquely upwards towards the aerial from ground level. Vertically-polarized ignition interference might be received from traffic passing outside a house in this way.

Coaxial Feeder, Vertically-Polarized Aerial.—There will be some sensitivity to horizontal polarization owing to reception on the horizontal part of the feeder. Reception of vertically-polarized waves will occur on the vertical part of the feeder as a result of both effects described in Sections 2.2 and 2.3. One consequence is that the vertical radiation pattern of the aerial will be different in different directions. It might be expected to include lobes directed obliquely downwards.

The performance of the aerial will depend upon whether the inner conductor of the feeder is connected to the upper or the lower limbs of the dipole. In the domestic radio trade it is usual to connect the inner conductor to the upper limb, as shown in Fig. 1(*a*). The reason advanced is that the lower limb, being nearer the ground, is nearer to earth potential and is therefore better connected to the outer conductor. This argument is far from convincing, but it is only fair to point out that the usual method of connection has some justification when the vertical part of the feeder is very close to the lower limb of the dipole. In the limiting case illustrated in Fig. 3(*a*) the aerial behaves like

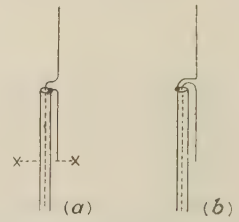


Fig. 3.—Vertical dipole with lower limb very close to coaxial feeder.

(*a*) Lower limb connected to outer conductor.
(*b*) Upper limb connected to outer conductor.

a sleeve dipole, so that in a very narrow band of frequencies the feeder is open-circuited at XX insofar as currents on the outside are concerned. This arrangement would be unsatisfactory for television, because the bandwidth in which the impedance would be sufficiently constant, and in which reception due to the outer conductor of the cable would be sufficiently small, would be inadequate. The reverse arrangement shown in Fig. 3(*b*) would be even less satisfactory. The radiation resistance would be very low (about 13 ohms) and the aerial would be particularly sensitive to currents on the outside of the feeder.

The fact that reception on the outside of the feeder can be reduced to small proportions in the arrangement shown in Fig. 3(*a*) implies that the contributions to the received signals from the two effects described in Sections 2.2 and 2.3 are equal and opposite. If the spacing between the dipole and the vertical part of the feeder is increased, as in Fig. 1(*a*), the cancellation will no longer be complete, and as the experiments described below confirm, it is completely lost when the spacing is $\lambda/8$.

(3) RADIATION-PATTERN MEASUREMENT

(3.1) General

An adequate representation of the 3-dimensional radiation pattern would be obtained by measuring radiation patterns in three perpendicular planes, i.e. one horizontal radiation pattern (h.r.p.) and two vertical radiation patterns (v.r.p.). It would be necessary to measure each of these three radiation patterns with each polarization in turn unless one polarization could be excluded by symmetry.

It is essential that the radiation patterns should be measured in such a way that the effect of the feeder as installed should be

clearly shown, but this is not easy. If measurements are made in the usual manner by rotating the aerial about a vertical axis, measurement of the v.r.p. entails running the feeder away horizontally for a sufficient distance and rotating it with the aerial. This would be inconvenient with a full-scale Band I aerial.

Strafford* has discussed the techniques used in the development of commercial receiving aerials, but has described only the measurement of the h.r.p. and that for the wanted polarization. Luscombe† measured the performance of a number of commercial receiving aerials, all directional and all designed for vertical polarization. He measured the h.r.p. in the usual way, as shown in Fig. 4(*a*). In order to obtain the v.r.p. he changed

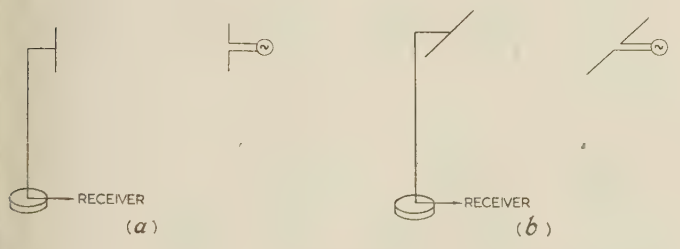


Fig. 4.—Methods of measuring radiation patterns of a vertical dipole.

(*a*) Measurement of horizontal radiation pattern.
(*b*) Usual method of measuring vertical radiation pattern (effect of feeder not correctly determined).

the aerial under test to horizontal polarization and again measured the h.r.p., as shown in Fig. 4(*b*). The fixed transmitting aerial was oriented for vertical polarization when the h.r.p. measurement was made, and for horizontal polarization when the v.r.p. measurement was made, so that only the wanted polarization was taken into account. Luscombe pointed out that, in using the arrangement shown in Fig. 4(*b*) for the v.r.p. measurement, the effect of the feeder was not correctly assessed, and he attributed discrepancies between back-to-front measurements to this cause. He suggested that more consistent results would have been obtained had a balanced feeder been used rather than a coaxial feeder.

(3.2) Measurements to Determine the Effect of the Feeder

The investigation was primarily concerned with aerials for frequencies between 41 and 216 Mc/s, but measurements were made in the neighbourhood of 450 Mc/s, using small-scale models.

The h.r.p.'s were measured in the usual manner, as shown in Fig. 4(*a*), and in order to measure the v.r.p.'s the arrangement shown in Fig. 5 was adopted. The section of cable AB represented part of the vertical feeder used in practice, its length being about 2λ. Some comparison measurements were made with lengths of $2{\cdot}25$ and $2{\cdot}5\lambda$, but this change had a negligible effect on the measured patterns. Thus the results may be applied, at least qualitatively, to aerials with longer feeders. Currents on the vertical feeder section BC would not contribute to the result, since the transmitting aerial was horizontal, and the horizontal feeder section CD would have little effect owing to its proximity to the ground. No attempt was made to simulate, in the model, the effect of reflection by the ground (see Section 2.3).

The feeder used was a $\frac{1}{4}$in-diameter 70-ohm polythene-insulated cable bonded to a metal support giving an overall effective diameter of about 8 mm ($0{\cdot}012\lambda$). Although the diameter of the supporting pole of a receiving aerial is not normally greater than this figure, it may be smaller but it will not be less than $0{\cdot}005\lambda$. Thus the effect of the type of support

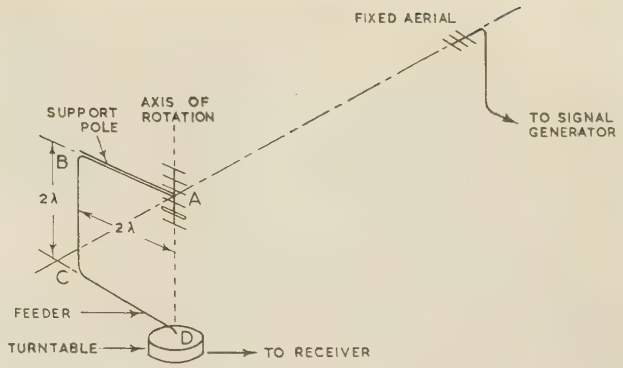


Fig. 5.—Method of measuring vertical radiation patterns.

and feeder normally used may be somewhat less than indicated by these measurements. However, the current induced on the feeder decreases slowly with diameter; a change from $0{\cdot}012$ to $0{\cdot}005\lambda$ should reduce the current by about 20% only.

(3.2.1) Measurements with a Simple Dipole.

The first set of measurements was made with a simple vertical dipole $0{\cdot}46\lambda$ long at 450 Mc/s, spaced $\lambda/8$ from its feeder. The dimensions of this model were chosen to represent a full-scale aerial used in band I (41–68 Mc/s), but the results should provide at least a qualitative guide to the performance of dipoles used at the higher frequencies. Measurements were made at 400, 450 and 500 Mc/s. The set of patterns for 450 Mc/s is shown in Fig. 6; similar results were obtained at the other frequencies.

If no current were excited on the outer conductor of the feeder, the h.r.p.'s would all be circular and the v.r.p.'s would be symmetrical about the horizontal and vertical axes. For the balanced dipole, where the outer conductor is excited only as a

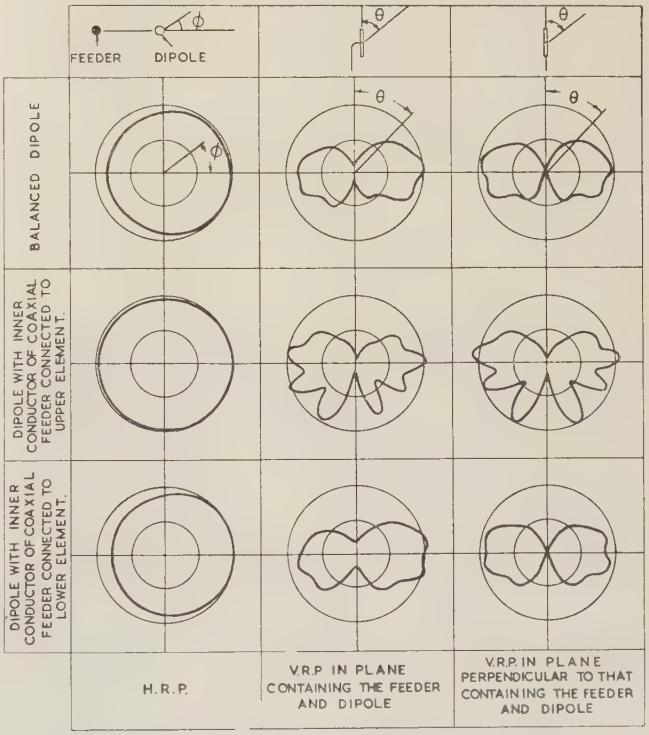


Fig. 6.—Measured radiation patterns of a $\lambda/2$ dipole.

The larger and smaller circles represent relative field strengths of 1·0 and 0·5 respectively.

* Strafford, F. R. W.: 'Receiving Aerials for British Television', *Proceedings I.E.E.*, Paper No. 1287 R, April, 1952 (99, Part IIIA, p. 631).
† Luscombe, G. W.: 'Television Aerials', *Wireless Engineer*, 1953, 30, p. 82.

parasitic element, the distortion is fairly small. When no balance/unbalance transformer is used, so that the outer conductor is excited directly as a driven element, the distortion is much greater. Lobes pointing obliquely downwards at about 40° to the vertical are symptomatic of travelling waves on the feeder. On the whole, the effect of the feeder was more marked when the inner conductor was connected to the upper limb of the dipole, according to the usual practice in the trade, than when it was connected to the lower limb.

(3.2.2) **Measurements with a Directional Aerial.**

Experiments were performed at 455 Mc/s with a vertically-polarized 6-element Yagi aerial having a metallic cross-arm to which the elements were bonded. This type of aerial was chosen as being typical of the more directional type of domestic receiving aerial in general use in band III. The feeder was attached to the supporting pole, which was parallel to the radiating elements. Thus the pole and the feeder behaved jointly as a parasitic element forming an integral part of the aerial system. A feature common to most Yagi aerials is their low radiation resistance. Consequently the radiator is usually folded, in order to raise the input impedance and enable it to be matched to the feeder. This practice was followed in the design of the aerial used in the experiments, since, as will appear, folding influences the effect of an unbalanced feeder. Of many known types of balance/unbalance transformers, the one selected made use of the radiator itself, the coaxial cable being taken to the feed point inside one limb of the dipole, as shown in Fig. 7(*a*).

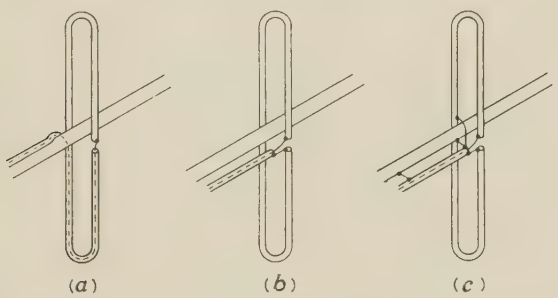


Fig. 7.—Methods of connecting a coaxial cable to a folded dipole.

(*a*) Hollow dipole used as balance/unbalance transformer.
(*b*) Direct connection.
(*c*) Outer conductor bonded to mid-point of dipole, cross-arm and supporting pole.

The dipole itself then behaves in a manner similar to the well-known 'Pawsey stub'.

When a balance/unbalance transformer is not used there are two possible methods of connecting a coaxial cable directly to a folded dipole. In the first [Fig. 7(*b*)], the inner and outer conductors are connected to the dipole terminals, but the outer conductor is insulated from all other parts of the aerial; this is the normal commercial practice. In the other method the outer conductor of the feeder is bonded at intervals to the supporting pole and to the cross-arm of the aerial, as shown in Fig. 7(*c*). Measurements were made with both methods of connection.

Figs. 8 and 9 show sets of measured radiation patterns. Each horizontal row corresponds to one particular aerial arrangement or method of connection, and the patterns are correctly scaled relative to each other. In Fig. 8, the upper row shows the patterns obtained when a balance/unbalance transformation was effected by the method shown in Fig. 7(*a*). The effect of connecting the feeder directly to the aerial, as in Fig. 7(*b*), with the inner conductor of the feeder connected in turn to the upper and lower halves of the driven element, is shown in the other two rows of Fig. 8. Reversing these connections reverses the phase

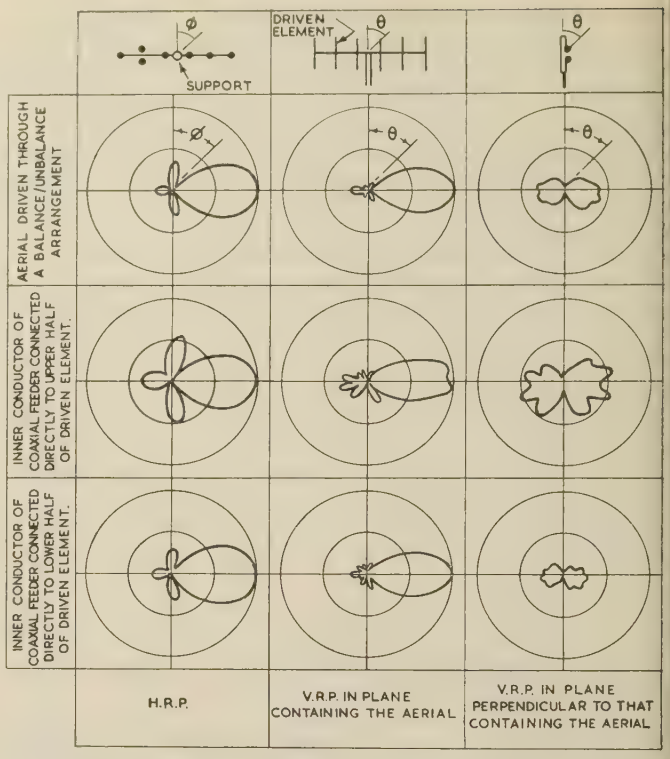


Fig. 8.—Measured radiation patterns of a Yagi aerial.

Driven element connected by methods shown in Figs. 7(*a*) and (*b*).
The larger and smaller circles represent relative field strengths of 1·0 and 0·5 respectively.

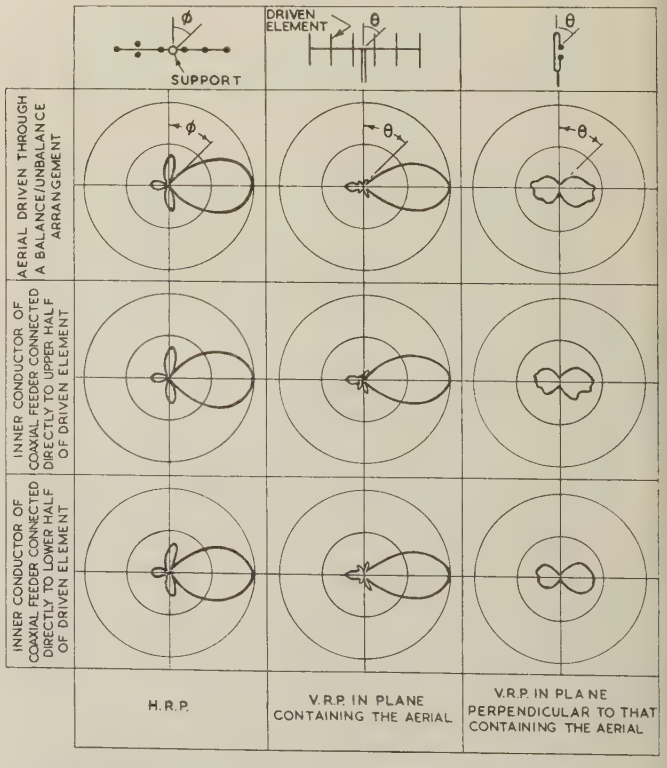


Fig. 9.—Measured radiation patterns of a Yagi aerial.

Driven element connected by methods shown in Figs. 7(*a*) and (*c*).
The larger and smaller circles represent relative field strengths of 1·0 and 0·5 respectively.

of the wanted component of the signal and that of the component associated with reradiation from the feeder (acting as a parasitic element). However, it leaves unchanged the phase of that component of the signal due to the lack of a balance/unbalance transformer (the feeder acting as a driven element).

The results shown in Fig. 8 suggest that the effects of the feeder acting as a driven element and as a parasitic element respectively tend to reinforce one another when the inner conductor is connected to the upper limb of the dipole, and to oppose one another with the reverse method of connection. In the former case, the lobes directed obliquely downward, which would be expected as a consequence of travelling waves on the feeder, are plainly visible in both vertical radiation patterns.

When the feeder was bonded at intervals to the cross-arm and supporting pole, as in Fig. 7(*c*), the results shown in Fig. 9 were obtained. It is apparent that the distortion of the radiation patterns in the absence of a balance/unbalance transformer has been greatly reduced. This effect may be explained theoretically, ignoring the effect of the metallic cross-arm and pole, by an extension of the procedure outlined in Section 2.3. It is found that the component of the received signal due to the lack of a balance/unbalance transformer is halved by bonding the outer conductor of the feeder to the mid-point of the folded dipole. The radiation patterns shown in Fig. 9 indicate that this result remains at least qualitatively true when the pole and cross-arm are metallic. Additional experiments were performed in order to determine to what extent, if any, the improvement had been caused by bonding between the outer conductor of the feeder, cross-arm and pole at points other than the feed point. With the outer conductor bonded only at the point nearest the dipole terminals, results intermediate between those shown in Figs. 8 and 9 were obtained. When it was bonded at four points in the first half-wavelength of the feeder, the patterns did not differ significantly from those shown in Fig. 9.

(4) CONCLUSIONS AND PRACTICAL IMPLICATIONS

The investigation described was limited in scope, and it would be unwise to apply the results in a quantitative sense to all types of aerial used for domestic reception. The principal conclusion to be drawn is that the effect of connecting a balanced aerial directly to a coaxial cable ought not to be ignored. Before completing the design of an aerial to be used in this way, measurements should be made to determine whether and to what extent currents on the feeder will impair the performance.

If a balanced feeder is connected to a balanced aerial, or if a coaxial feeder is connected via a balance/unbalance transformer, the feeder will be excited only as a parasitic element. In this condition its effect is small, even when the aerial is vertical. It would, of course, have no effect on a horizontal aerial.

If a coaxial cable is connected directly to a balanced dipole, the v.r.p. of the latter is seriously distorted, owing to the direct excitation of currents on the outer of the cable. One feature of this effect is a sensitivity to vertically-polarized interference generated near to and below the aerial, e.g. ignition interference from a road outside the house on which the aerial is mounted. It is believed that this would still be true in the case of a horizontal dipole, so that the effect may sometimes contribute to the reception of ignition interference in band II. The effect on the performance of the Yagi aerial investigated was much smaller. It may be that the higher gain of the aerial makes reception by currents on the outside of the feeder less important, but it is also possible that the metallic cross-arm and supporting pole are important in this connection. It would therefore be unwise to apply the conclusions reached to a similar aerial with a non-metallic cross-arm and pole.

When a folded dipole is to be connected directly to a coaxial cable without a balance/unbalance transformer, it is desirable to bond the outer conductor of the cable to the mid-point of the folded dipole. If the dipole is supported on a metal rod, such as the cross-arm of a Yagi aerial, it is also advisable to bond the first half-wavelength of the feeder to this rod at intervals. With the aerial investigated the last-mentioned procedure rendered the use of a balance/unbalance transformer unnecessary.

It is nevertheless preferable to avoid a direct connection between a balanced aerial and a coaxial cable. Assuming the appropriate receiver input in each case, two alternative arrangements appear worthy of consideration—a coaxial cable and some kind of balance/unbalance transformer at the aerial, or a balanced twin feeder connected directly to the aerial. The latter may be either screened or unscreened.

Coaxial and screened twin cables have the advantage of being unaffected by the weather and by proximity to lossy materials such as walls, and they are unaffected by painting. The provision of a balance/unbalance transformer for use with a coaxial cable would add to the complexity of the aerial, particularly in band I and to a lesser extent at higher frequencies. A simpler arrangement using screened twin cable involves a somewhat higher loss because of the higher attenuation in this feeder.

The use of unscreened twin feeder, insulated with black polythene, offers a simpler solution, since no balance/unbalance transformer is required. There seems to be little justification for the belief that unscreened feeder is more susceptible to local interference, provided that its balance is not impaired. Flat forms of feeder are, however, subject to a considerable increase in loss when exposed to rain. A tubular form is somewhat better, but may compare unfavourably with the coaxial cables in common use when wet. It seems likely that the tubular form will also confer some immunity to the effect of dirt and paint, but this has not been established. The attenuation and the balance of this cable are affected greatly by proximity to walls, etc., but this effect can be mitigated by the use of spacers. Since the characteristic impedance of tubular twin feeder is about 300 ohms, mismatch loss would be somewhat higher unless special precautions were taken.

The conclusions stated above apply only when the receiver has a well-designed input circuit, which could take any of three forms. It could be screened and unbalanced (this implies that signals picked up on a single wire connected to the earthy terminal would not be received); it could be balanced, with or without an earthed centre tap; finally, it would be possible—although perhaps less easy—to replace either of the other two types of circuit with a 'floating' input circuit, such as a primary winding of very low capacitance. Unfortunately, the input circuits of domestic receivers do not usually fall into any of these classes, so that push-push currents may contribute to the received signal whether a coaxial or a balanced feeder is used. Until receivers are improved in this respect an ideal solution does not exist.

(5) ACKNOWLEDGMENTS

The measurements described were performed by Mr. K. Hughes and Mr. I. D. B. Millar, and their assistance is gratefully acknowledged. The authors are indebted to the Chief Engineer of the British Broadcasting Corporation for permission to publish the paper.

621.372.5

The Institution of Electrical Engineers
Paper No. 3114 E
Jan. 196(

©

CLASSES OF 4-POLE NETWORKS HAVING NON-LINEAR TRANSFER CHARACTERISTICS BUT LINEAR ITERATIVE IMPEDANCES

By Professor E. COLIN CHERRY, D.Sc.(Eng.), Associate Member.

(*The paper was first received 22nd April, and in revised form 6th July*, 1959.)

SUMMARY

The conventional representation of an electric circuit as a pattern of branches, nodes and meshes (topological graph) is historic. The less familiar representation as a set of contiguous rectangles, proposed by the author[1] in 1951 from a suggestion made by Hering[2] (1927), has many conceptual advantages;* not only does it represent the network pattern topologically but it displays other physical properties: the dual of the circuit is included; the element energies and co-energies (energy duals) are shown; current and voltage magnitudes are represented, etc. This rectangle representation of a planar circuit is briefly reviewed and its special application to non-linear circuits is explained.

From simple geometrical symmetry of the rectangle diagrams it becomes immediately obvious that iterative structures (lattices, bridged-T, etc.) may be constructed from dual non-linear resistive elements having linear iterative impedances. If these can be constructed practically, they might be connected in cascade, without mutual interaction, as non-linear equalizers analogous to Zobel's constant-resistance phase-equalizers.

The analogy to Zobel's networks is shown to be surprisingly complete, though the author can yet find no physical relation between these and the present non-linear networks.

(1) THE RECTANGLE-DIAGRAM REPRESENTATION OF A PLANAR CIRCUIT

The representation of a planar electric circuit, both topologically and physically, by a set of contiguous rectangles has been fully explained elsewhere;[1] it may be understood from the simple example of Fig. 1.

Fig. 1(*a*) shows a bridge circuit, driven by a generator $V(t)$ supplying a current $I(t)$; the bridge consists of linear resistances

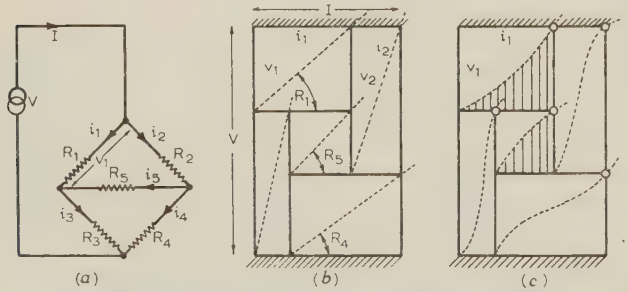


Fig. 1.—The rectangle representation of a planar circuit (in this example a bridge circuit).

R_{1-5} carrying potential-drops v_{1-5} and currents i_{1-5}. The same circuit is represented by the diagram of Fig. 1(*b*), as we may see thus: each rectangle represents one resistance, its sides having lengths i_r and v_r; the fact that, placed together, they fit into the overall rectangle of sides I and V follows from Kirchhoff's laws. Note, further, the aspect ratio of any rectangle, v_r/i_r, gives the resistance R_r, represented by the slope of the diagonal, as shown on every rectangle in Fig. 1(*b*). In such a diagram, all horizontal lines represent nodes and all vertical lines represent meshes.

Note again, if such a diagram as Fig. 1(*b*) be turned on its side, currents and voltages are interchanged everywhere; also, vertical (mesh) and horizontal (node) boundaries are exchanged. On its side, therefore, the diagram represents the dual of the original circuit of Fig. 1 (it happens to be self-dual in this example).

Note, further, that Kirchhoff's laws $\Sigma v = 0$ and $\Sigma i = 0$ are inherent numerically in these diagrams, as is also the principle of the conservation of energy (since by inspection $\Sigma v_r i_r = VI$).

(1.1) Application to Non-Linear Circuits

Fig. 2 illustrates a typical non-linear resistor characteristic; the ringed point, P, is the instantaneous working-point. It

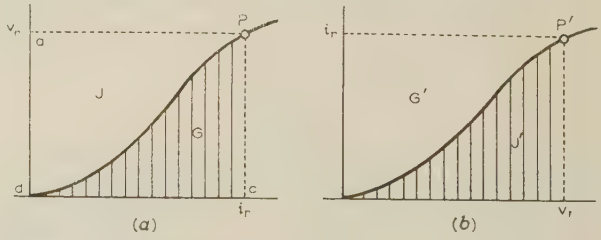


Fig. 2.—Non-linear resistance characteristic and its dual.

becomes clear immediately that such characteristics, enclosed within the rectangle *aPcd* may be fitted together into rectangle representations, as in Fig. 1(*c*), which shows the circuit of Fig. 1(*a*) when all the resistors have non-linear characteristics. This fact depends only upon Kirchhoff's laws.

In Fig. 1(*c*) the instantaneous working-points of each non-linear resistor have been ringed (top right-hand corner). Hence, as time varies and currents and voltages vary, these working-points slide up and down the characteristics, but always in such a way that the rectangles fit together within the overall generator rectangle $V(t)I(t)$, which, of course, varies in size as $V(t)$ and $I(t)$ vary.

(1.2) Division of Power into Content and Co-Content

Consider the rectangle in Fig. 2(*a*). The area lying below the characteristic, terminated at the working-point P, is called the content, and the area above is called the co-content[1]:

$$G = \int_0^{i_r} v\,di \quad \ldots\ldots\ldots \quad (1)$$

$$J = \int_0^{v_r} i\,dv \quad \ldots\ldots\ldots \quad (2)$$

* The idea originally arose from the theory of squared-squares, a classical problem in geometry. See, for example, Brooks, Smith, Stone and Tutte: *Duke Mathematical Journal*, December, 1940, 7.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

Prof. Cherry is Professor of Telecommunication, Imperial College of Science and Technology, University of London.

Then $G + J = P$, the instantaneous power in the non-linear element. Note that with linear elements, $G = J$, a degeneracy which had hidden certain fundamental properties of non-linear circuits until exposed by Millar.[1]

Fig. 2(*b*) shows the dual non-linear resistor. Here, v and i are interchanged. If the corresponding content and co-content are called G' and J', then, in this dual

$$G' = J$$
$$J' = G$$

or, perhaps, with any assigned constant of proportionality.

The author has shown elsewhere that rather similar duality rules may be established for non-linear inductors and capacitors.[1] Whether any such dual elements can be constructed physically is a practical question; they 'exist' conceptually and may be defined, as here, and are therefore available as a tool in non-linear circuit theory.

(2) SYMMETRY PROPERTIES OF RECTANGLE DIAGRAMS

It is now possible, by inspection, to see that rectangle diagrams may be constructed, using (*a*) linear and (*b*) dual non-linear resistors, in such ways that driving-point resistances remain linear.

(2.1) The Bridge or Lattice Structure (Symmetric)

Perhaps the simplest structure to visualize is the bridge or lattice. Fig. 3(*a*) shows a symmetric lattice section, constructed

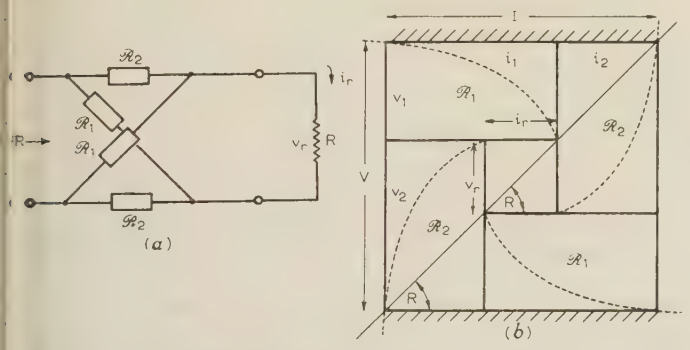


Fig. 3.—A symmetric bridge or lattice of non-linear resistances, having linear iterative resistances.

of dual non-linear resistances $\mathscr{R}_1$ and $\mathscr{R}_2$ with linear resistive termination R. This section then, has also a linear driving-point resistance R, as we may see from its rectangle representation, Fig. 3(*b*), which has complete rotational symmetry for all positions of the working-points on the curved i/v characteristics. Then if the boundaries I and V are normalized to be square, the inner rectangle $i_r v_r$, also becomes square; this normalization makes $R = 1$ and is used throughout.

This form of symmetric non-linear lattice circuit has been described[3] in British Patents Nos. 455189 and 457236. It is the particular form of presentation here which concerns us, since visual inspection enables us to see other forms of symmetry, leading to other networks.

(2.2) The Bridge or Lattice (Unsymmetric)

The rectangle representation of Fig. 4(*b*) has a simple diagonal symmetry and represents the asymmetric lattice of Fig. 4(*a*). For all working-points, the common corner (ringed) of the two constant-resistance squares, R, will be constrained so as to lie on the diagonal shown.

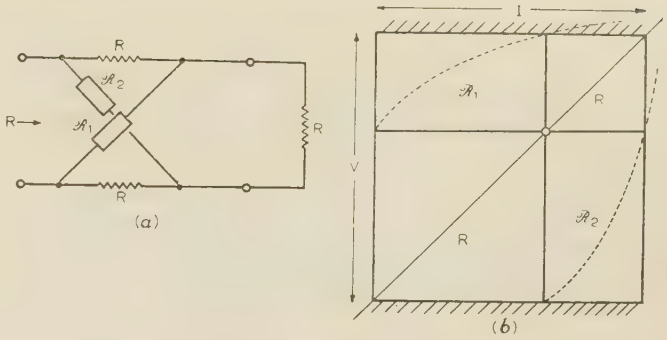


Fig. 4.—An asymmetric lattice.

(3) CALCULATION OF NON-LINEAR TRANSFER CHARACTERISTICS

The non-linear transfer characteristics of the lattices described may readily be determined algebraically if the non-linear resistor characteristics are given. Let these two dual characteristics be

$$i = \mathrm{F}(v);\ v = \mathrm{F}(i) \quad . \quad . \quad . \quad . \quad . \quad (4)$$

respectively. Then from the example of Fig. 3(*b*), we see the instantaneous relationships

$$\left.\begin{aligned} I &= i_1 + i_2 \\ V &= v_1 + v_2 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

$$i_2 = v_1/R \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

On the output terminals (central square) we have

$$v_r = V - 2v_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

Hence, substituting $v = v_1$ and $i = i_1$ in eqn. (4), we have

$$v_1 = \mathrm{F}(i_1) = \mathrm{F}(I - i_2) = \mathrm{F}\left(\frac{V - v_1}{R}\right) \quad . \quad . \quad (8)$$

Solving this equation for v_1 and writing these solutions as

$$v_1 = S(V) \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

we may substitute in eqn. (7) to give the non-linear voltage transfer characteristics as

$$v_r = V - 2S(V) \quad . \quad . \quad . \quad . \quad . \quad (10)$$

If the non-linear element characteristics be simple and single valued [such as the simple parabolic curves illustrated in Fig. 3(*b*)], eqn. (9) will give only a single usable solution. It should be noted, however, that even if one of the non-linear resistors $\mathscr{R}_1$ gives a voltage drop which is a single-valued function of current, its dual may require a multi-valued function (for example, Fig. 5).

It serves our purpose better here, especially in such multi-valued cases, to stress that an alternative to such algebraic

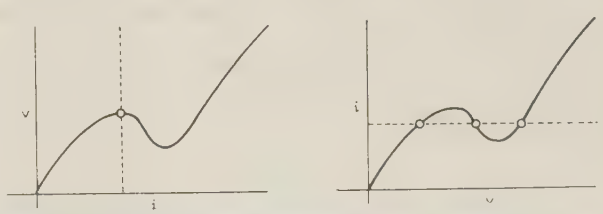


Fig. 5.—Dual non-linear resistance characteristics, one being multi-valued as a function of current.

computation lies in geometrical construction of the rectangle diagrams. To construct these, the length V may be assigned a set of suitable values, giving corresponding values of I, and so, a set of bounding rectangles. Then templates may be cut out, corresponding to the dual characteristics, and fitted in rotational symmetry, as in Fig. 3(*b*). Then the central square, $v_r i_r$ is obtained.

Such a use of templates may well provide a practically useful trial-and-error technique for solving non-linear problems of kinds other than those we are discussing, used in conjunction with these rectangle representations of circuits.

As a simple example, a parabolic characteristic and its dual,

$$i = v^2 \quad v = i^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (11)$$

have been used for $\mathscr{R}_1$ and $\mathscr{R}_2$ in the bridge (lattice) circuit of Fig. 3, and the input/output transfer ratio, V/v_r, determined. Illustrated in Fig. 6, this curve shows the balance point, $v_r = 0$, of this non-linear bridge.

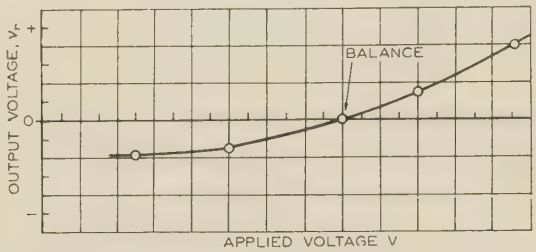


Fig. 6.—The transfer-ratio characteristic for a non-linear bridge using dual arms having parabolic voltage/current characteristics.

(4) LADDER NETWORKS

We may now show that classes of ladder networks and bridged-T networks exist which also have linear iterative impedances, though constructed of dual non-linear resistances.

There is a certain analogy here to Zobel's constant-resistance equalizers, which will be apparent.[4] Such networks consist of lattice, bridged-T and ladder forms (mutually transformable structures) constructed of conventional L and C dual elements and resistances; they have pure resistive iterative impedances and so may be connected in cascade, without mutual interaction, and used for phase correction. Our present circuits have similar circuit structures, using non-linear dual resistances, and could possibly be used for the correction of non-linearity.

(4.1) Simple Ladder Structures

The 2-terminal network shown in Fig. 7(*a*), constructed with dual non-linear resistances $\mathscr{R}_1$ and $\mathscr{R}_2$, has a constant driving-point resistance, R. This may be shown from its rectangle representation in Fig. 7(*b*), which is seen to have diagonal symmetry.

Turning the diagram on its side immediately shows the network to be self-dual. But if the topological dual of the circuit in Fig. 7(*a*) be formed, that of Fig. 7(*c*) results; however, the points marked * must be at the same potential and so may be joined, giving self-duality.

Note that, in this dual, the linear resistances, together with the overall linear resistance, have the dual values $1/R$, although they have been normalized to unit value here.

(4.2) A Simple Ladder

The circuit in Fig. 7(*a*) may be redrawn as a ladder half-section [Fig. 7(*d*)]. Such sections may be connected iteratively in cascade, since both the driving-point and termination resistances,

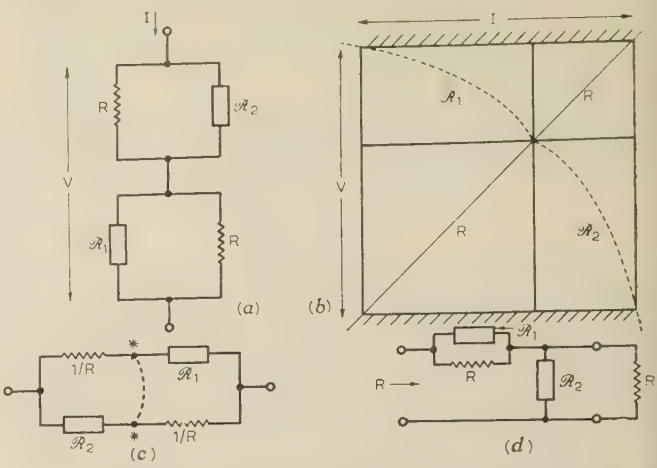


Fig. 7.—Constant-resistance arms, constructed with dual non-linear resistances.

R, are linear, at least for the forward (left to right) direction shown here.

As regards transmission in the backward (right to left) direction, we see that this ladder does not have a linear iterative resistance, since it would be driven and terminated, as shown for one section in Fig. 8(*a*). The driving-point resistance R^*

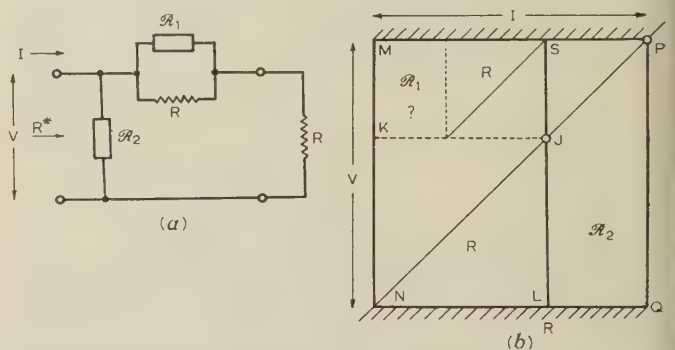


Fig. 8.—A reversed L-section has a driving-point resistance $R^* \neq R$.

cannot, in general, be linear, for if $R = R^*$ the circuit rectangle diagram would be bounded by V and I as in Fig. 8(*b*), with the working-point P (ringed) sliding on the straight diagonal as shown. Then, we may fit $\mathscr{R}_2$ into the rectangle diagram, but this supports the whole voltage V and may be represented by the rectangle PQRS. Next, the load resistor R represented by rectangle JKLN, being linear ($R = V/I$), must also have its working-point J sliding on the diagonal. This leaves the rectangle JKMS into which we must fit the rectangles representing $\mathscr{R}_1$ and the paralleled R. Clearly it cannot be done, since the (dual) rectangle for $\mathscr{R}_1$ must have a size and shape related at every working-point to $\mathscr{R}_2$.

Consequently, ladder networks constructed of such L-sections, as in Fig. 7(*d*), have constant iterative resistances only when driven mid-series and terminated mid-shunt.

We shall show in Section 6 that slightly more complicated ladder structures, consisting of symmetric T-sections, may be found which do, in fact, have linear iterative impedances for propagation in either direction.

(5) BRIDGED-T STRUCTURES

A restricted form of bridged-T structure, having linear iterative resistive impedances R for both directions of transmission, may be constructed starting from the section shown in Fig. 7.

(5.1) Degenerate Forms of Bridged-T Network with Linear Bilateral Iterative Impedances

The structure shown in Fig. 7(*c*) was seen to have constant resistance $1/R$, whilst the points marked * * are always at the same potential. If we connect a linear resistance, itself equal to $1/R$, between the points * *, as in Fig. 9(*a*), it will carry no

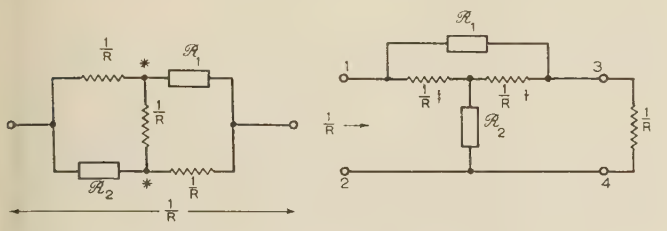


Fig. 9.—Degenerate bridged-T constant-resistance section.

current. The whole structure may then be redrawn as in Fig. 9(*b*), forming a bridged-T section; for propagation from left to right no current flows in the element $1/R$ marked †.

Now this section is completely symmetrical. Consequently, we could also propagate from right to left, with a linear termination $1/R$ connected across terminals 1 and 2; in this case no current would flow in the resistor $1/R$ marked ‡. Thus, reciprocity is applicable with respect to the two pairs of terminals 1, 2 and 3, 4.

(5.2) A More General Reciprocal Bridged-T Structure

One naturally wonders whether more general bridged-T structures exist, with linear resistive iterative impedances, for bilateral transmission, in which perhaps current flows in all three series arms. Now we have already observed some analogy to Zobel's constant-resistance (equalizer) networks,[4] though no physical reason for the analogy is apparent to the author. Let us, however, take the hint and try the structure shown in Fig. 10, which is analogous to one of the Zobel§ structures.

We see that this structure may be produced from that of Fig. 9 as follows: if we add any linear resistance in series with $\mathscr{R}_2$ and the dual of this resistance in parallel with $\mathscr{R}_1$, the resulting series and parallel combinations are still duals and the overall constant-resistance properties of Fig. 9 are unchanged. The network now contains a delta connection of linear resistances; transform these to the star connection and the result is the circuit of Fig. 10.

Since we are dealing with non-linear systems a more formal proof may be called for. It is illuminating to use our rectangle diagram approach.

Fig. 10 also shows the rectangle diagram. In this, V and I have been normalized so that the overall rectangle becomes a square; again the linear termination R is always square ($R = 1$) and its working-point (ringed) slides on the main diagonal as shown. Another condition is that, since $\mathscr{R}_1$ and $\mathscr{R}_2$ are dual at all working currents, their rectangles are always identical; their side lengths have been marked a and b. From the geometry we may mark other lengths to have the values shown on the diagram. The larger rectangle corresponding to one linear resistance R/c (where c is an arbitrary constant) is shown having sides a and $a + d$; its working point slides on its own diagonal, as shown. That corresponding to the other R/c resistance is smaller but must have the same side ratio.

We have only to determine whether the resistance marked R^* is linear or not; this depends upon whether the ratio of its rectangle sides remains constant for all working points; one side is a, and the other is called x, which is yet undetermined.

§ In Zobel's constant-resistance equalizers, $\mathscr{R}_1$ and $\mathscr{R}_2$ are replaced by dual linear L and C elements, or by dual combinations of them.

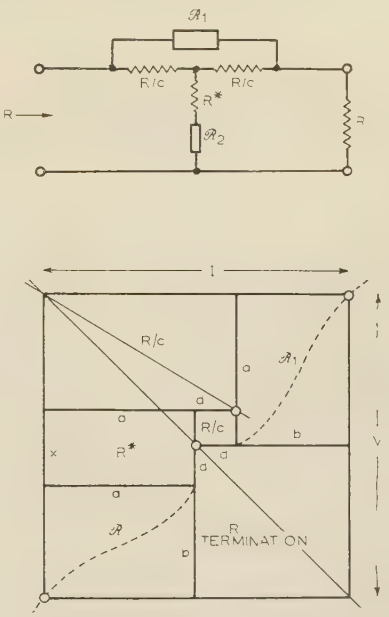


Fig. 10.—A class of bridged-T constant-resista nce sections.

$$R^* = \frac{c^2 - 1}{2c} R \qquad c = \text{A constant.}$$

We have scaled V and I for all working values. Hence, from the similarity of the two R/c rectangles, we may equate the ratios of their sides:

$$(x - d)/d = \frac{1}{c} \quad . \quad . \quad . \quad . \quad (12)$$

$$(a - x + d)/(a + d) = \frac{1}{c} \quad . \quad . \quad . \quad . \quad (13)$$

Now the resistance under question, R^*, is given by the ratio x/a. From eqn. (12), $x = d(1 + c)/c$; substitute into eqn. (13), giving $a = 2d/(c - 1)$. Then the resistance R^* has the normalized magnitude

$$\frac{x}{a} = \frac{(c^2 - 1)}{2c} \quad . \quad . \quad . \quad . \quad . \quad (14)$$

which is a constant, since c is a constant. Removing the normalizing, the actual value of R^* is then

$$R^* = \frac{c^2 - 1}{2c} R$$

This parameter, $(c^2 - 1)/2c$, is exactly the same parameter in the same circuit of Zobel's equalizer.[4] But the physical relation between Zobel's circuits, which use LC linear dual elements, and ours here, which use non-linear dual resistances, is not clear. It may be fortuitous or there may be some deeper topological reason.

(5.3) The Dual Bridged-T Structure

If the rectangle diagram of Fig. 10 be turned on its side, voltage and current are everywhere interchanged, and the dual network results. The linear elements dualize in the conventional way, whilst nodes and meshes are interchanged. The non-linear elements $\mathscr{R}_1$ and $\mathscr{R}_2$ dualize into one another.

Fig. 11 shows the corresponding dual circuit which again has linear iterative resistive impedances for both directions of transmission.

Once again we find the structure identical with one of Zobel's linear constant-resistance equalizers.

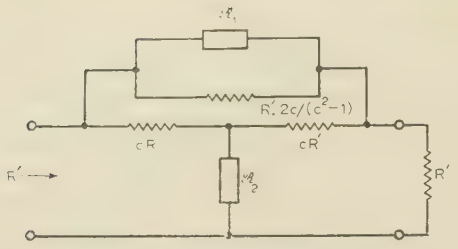


Fig. 11.—The dual structure to that of Fig. 10.
R' is dual to R in Fig. 10 and so has the value $1/R$.

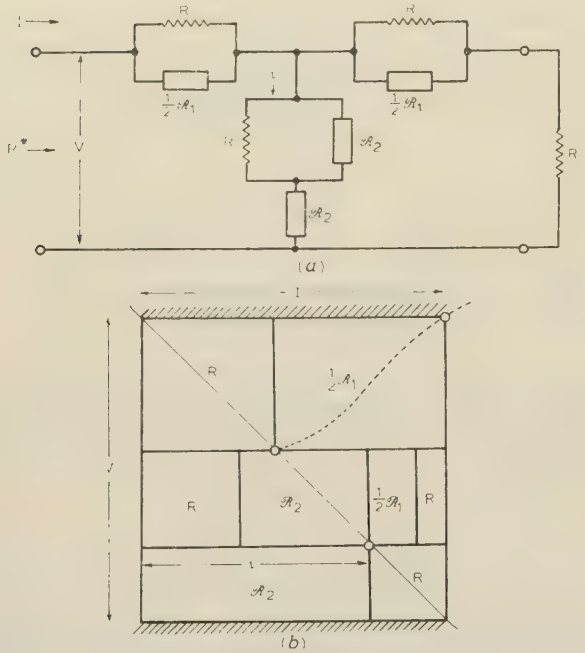


Fig. 12.—A Zobel-type ladder structure for which the input resistance R^* cannot, in general, be linear.

(6) A COMPOUND LADDER STRUCTURE, WITH LINEAR ITERATIVE IMPEDANCES?

This striking, but unexplained, analogy to Zobel's networks has not yet been followed out to the end. There remains one more structure which it is tempting to try experimentally in the present context. It is shown in Fig. 12(*a*); in Zobel's case,[4] the dual non-linear resistances $\mathscr{R}_1$ and $\mathscr{R}_2$ are replaced by dual linear L and C elements.

Fig. 12(*b*) shows the rectangle diagram. We see at once that the input resistance V/I cannot be linear except by fortuitous choice of characteristics. For one thing, the current i which feeds both R and $\mathscr{R}_2$ in parallel enters $\mathscr{R}_2$ subsequently; hence, both these non-linear elements, $\mathscr{R}_2$, cannot be operating at the same working-points. In no case, then, can the two rectangles $\mathscr{R}_2$ in Fig. 12(*b*) be identical at all working currents.

In this case therefore, $R^* \neq R$, and the analogy fails.

(7) THE RECIPROCITY THEOREM

Since no finding may be said to be trivial in non-linear theory, it may be worth mentioning that all the lattice, T and bridged-T structures dealt with here, possessing linear iterative impedances, obey the reciprocity theorem with respect to their input and output terminals. This is clearly so from the mirror-symmetry of the structures.

(8) REFERENCES

(1) CHERRY, E. C.: 'Some General Theorems for Non-linear Systems Possessing Reactance', *Philosophical Magazine*, 1951, **42**, p. 1161 (see also MILLAR, W.: 'Some General Theorems for Non-Linear Systems Possessing Resistance', *ibid.*, p. 1150).

(2) HERING, E.: 'Neue Methoden der graphischen Netzberechnung', *Elektro-Journal*, 1927, **7**, pp. 17, 51 and 98.

(3) SIEMENS UND HALSKE A.G.: British Patents Nos. 455189 and 457236.

(4) ZOBEL, O. J.: 'Distortion Correction in Electrical Circuits with Constant Resistance Recurrent Networks', *Bell System Technical Journal*, 1928, **7**, p. 438.

621.396.33: 621.391.812.3: 621.376.55

The Institution of Electrical Engineers
Paper No. 3125 E
Jan. 1960
©

PHASE-SHIFT KEYING IN FADING CHANNELS

By H. B. VOELCKER.

(*The paper was first received 4th April, and in revised form 30th June, 1959.*)

SUMMARY

Phase-shift keying (p.s.k.) is discussed as a modulation technique for transmitting digital data over radio circuits subject to fading. The modest bandwidth requirements of p.s.k. modulation suggest that it can not only alleviate spectrum crowding, but can also transmit traffic with fewer errors. The theoretical results presented here indicate, however, that the random phase perturbations inherent in fading radio signals cause unavoidable degradation in the performance of p.s.k. systems. Experimental data which partially support the theoretical results are cited, and comments relevant to the improvement of p.s.k. systems are included.

(1) INTRODUCTION

Long-range radio circuits have long been used for the transmission of teleprinter traffic and other forms of digital data. Marked improvements in the reliability of such circuits began to accrue about two decades ago, as frequency-shift keying (f.s.k.) gradually replaced amplitude modulation for digital data transmission. However, the steadily increasing crowding of the radio spectrum, together with demands for ever lower error rates, has fostered a continuing search for better modulation techniques. It was recognized long ago that phase-shift keying (p.s.k.) offers attractive possibilities for bandwidth conservation and for reducing the number of errors caused by white noise, but practical difficulties delayed its exploitation until quite recently.

In 1957, an interesting field test was conducted over a 5000-mile radio path to evaluate the performances of a modern f.s.k. system and a new p.s.k. system. While the operating procedures and many of the results of the test have already been reported,[1] a brief summary of this information is given in Section 9. Perhaps the most striking result was the comparative performance of the two systems. For relatively low values of signal/noise ratio, the p.s.k. system demonstrated a fairly constant superiority; at relatively high ratios, however, the f.s.k. system showed a superiority which increased steadily with the signal/noise ratio. This curious comparative performance occasioned the theoretical analysis and the comments set forth in the paper.

(2) PRELIMINARY DISCUSSION OF PHASE-SHIFT KEYING

The use of p.s.k. modulation for constant-rate data transmission requires that a stable carrier wave be available at the transmitter. At equally spaced instants in time, the carrier may be shifted in phase either 180°—the binary case—or $\pm 90°$, 180°—the quaternary case—to transmit either one or two bits of information. If quaternary keying is used, it is easy to show that the time-bandwidth product of the modulation is 1/2; this represents maximum channel loading if inter-symbol interference is to be avoided. (For simplicity, only a single phase-keyed carrier is considered in the paper; the extension to frequency-multiplexed systems is obvious. Furthermore, p.s.k. data pulses need not have a 'square' shape in the time domain, since the associated $\sin x/x$ spectrum is sometimes inconvenient. Costas, for example, has proposed a p.s.k. system which radiates signals with essentially 'square' spectral characteristics.[2])

Demodulation of p.s.k. signals requires a phase reference with which incoming pulses can be compared. Schemes for providing phase references fall into two classes, namely stable sources and transmitted references. If the transmission medium has constant phase characteristics, very stable frequency sources, e.g. caesium clocks, may be used for modulation and demodulation. After initial synchronization, the inherent stability of the sources should ensure reliable operation with only occasional adjustments to compensate for drift. Unfortunately, phase perturbations in most physical media require that transmitted phase references be used.

Schemes for transmitting phase references are many and varied, and the paper is concerned mainly with p.s.k. demodulation via 'differentially coherent detection'. With this technique, each signal pulse is detected by phase comparison with the (stored) preceding signal pulse. Information is thus conveyed by the phase differences between successive pulses, and redundant information need not be transmitted for reference purposes. Section 10 explains why differentially coherent detection is well suited for p.s.k. transmission over fading radio circuits, where phase perturbations are more or less continuous.

Before proceeding with an analysis of p.s.k. detection, it is worth while to review some of the statistical characteristics of fading radio waves.

(3) STATISTICAL CHARACTERISTICS OF FADING RADIO WAVES

Although the statistics of fading signals have received wide attention in the literature,[3–5] this brief review is included because the bivariate statistics which are important in p.s.k. studies are unfamiliar to many engineers. The mathematics are identical with those of noise theory, since it has proved convenient to view a fading signal as the envelope of narrow-band normally-distributed noise which may or may not contain a steady (specular) component.

Suppose that a continuous stable carrier signal is transmitted through a fading channel. At some instant, t, the received signal can be represented by the vector sum of two independent, quadrature vectors, x_t and y_t, each of which exhibits stationary normal statistics with zero mean value and average power σ_s^2. The probability density function (p.d.f.) of the received signal vector is therefore

$$\mathrm{p}(x_t, y_t) = \mathrm{p}(x_t)\mathrm{p}(y_t) = \frac{1}{2\pi\sigma_s^2} \exp\left(-\frac{x_t^2 + y_t^2}{2\sigma_s^2}\right) \quad . \quad (1)$$

By converting this expression to polar co-ordinates (E_t, θ_t) it is easy to show that the vector modulus, E_t, exhibits the well-known Rayleigh p.d.f. and that all values between $\pm\pi$ are equally probable for the phase, θ_t.[6, 7]

Ostensibly it would appear that p.s.k. modulation cannot be transmitted through a fading channel because of the random phase characteristics of the received wave. This is not true,

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. Voelcker is in the Department of Electrical Engineering, Imperial College of Science and Technology, University of London, and was formerly at the U.S. Army Signal Research and Development Laboratory, New Jersey, U.S.A.

however, for the preceding result tells nothing about the rapidity of phase changes; it merely shows that the phase of a single random sample is unpredictable. Temporal variations can be evaluated by describing $x_{t+\tau}$ and $y_{t+\tau}$, the quadrature components at time $t + \tau$, in terms of conditional p.d.f.'s involving x_t and y_t; these take the following form:[7]

$$p(V_{t+\tau}|V_t, \phi_\tau) = \frac{1}{\sigma_s\sqrt{[2\pi(1-\phi_\tau^2)]}} \exp\left[-\frac{(V_{t+\tau}-\phi_\tau V_t)^2}{2\sigma_s^2(1-\phi_\tau^2)}\right] \quad . \quad . \quad . \quad . \quad (2)$$

where $\phi_\tau = \text{Av}\,[V(t)V(t+\tau)]/\sigma_s^2$ and $0 \leqslant \phi_\tau \leqslant 1$.

A p.d.f. may now be written for the received signal at the two instants, t and $t + \tau$, in terms of x_t, y_t, $x_{t+\tau}$ and $y_{t+\tau}$. This expression can in turn be converted to polar co-ordinates (E_t, $E_{t+\tau}$, $\Delta\theta_\tau$), where $\Delta\theta_\tau$ is the random phase change over τ seconds. Finally, by integrating over all values of E_t and $E_{t+\tau}$, the following rather formidable p.d.f. can be obtained for $\Delta\theta_\tau$:

$$p(\Delta\theta_\tau|\phi_\tau) = \frac{(1-\phi_\tau^2)}{2\pi(1-\mu^2)}\left[1 + \frac{\mu}{\sqrt{(1-\mu^2)}}\left(\frac{\pi}{2} + \arcsin\mu\right)\right] \quad . \quad . \quad . \quad . \quad (3)$$

where $\mu = \phi_\tau \cos\Delta\theta_\tau$.

This expression was derived by Bramley in his investigation of space-diversity reception, although he was concerned with phase differences between aerials rather than with temporal phase shifts.[8] While the investigation of p.s.k. detection in later sections does not involve eqn. (3) directly, some discussion of eqn. (3) is worth while to illustrate certain natural constraints governing the performance of p.s.k. systems on fading circuits.

The key parameter in eqn. (3) is ϕ_τ, the correlation coefficient of each quadrature component over the interval τ. Price has reported experimental results which provide a good deal of physical insight into ϕ_τ.[9] To determine ϕ_τ for a short E-layer path he heterodyned the 5 Mc/s signal from WWV to 20 c/s with a very stable reference oscillator; this yielded a 20 c/s sine wave modulated by the noise-like envelope of one of the quadrature components of the received signal vector. He then autocorrelated the 20 c/s signal, and obtained ϕ_τ directly from the envelope of the autocorrelogram. His results show that the correlation function was roughly Gaussian in shape with a standard deviation in the order of 1 sec. More generally, his results agree with predictions based on the physics of turbulent media, in that ϕ_τ should be an even, monotonically decreasing function of τ.

A plot of eqn. (3) for Price's values of ϕ_τ is shown in Fig. 1. It is clear that when τ is very small and $\phi_\tau \simeq 1$, phase changes greater than a few degrees, say, are possible but improbable. Conversely, as τ increases and ϕ_τ decreases, large random excursions in phase become increasingly probable. Direct experimental measurements of temporal phase fluctuations have been reported recently by Brennan and Phillips for medium-frequency E-layer propagation,[10] and by Landmark and Hagfors for ionospheric scatter propagation.[11] These authors' results generally lend credence to eqn. (3) and the theory underlying it, and to the comments on the nature of ϕ_τ.

Certain conclusions can be drawn immediately from eqn. (3) regarding the performance of p.s.k. transmission in fading channels. The transmission of information requires time; thus, if τ is the basic pulse length of a p.s.k. system, eqn. (3) shows that errors will occur in binary or quaternary phase-keyed signals even though no other disturbances, e.g. additive noise, are present. This follows because the probabilities of large random phase excursions exceed zero for $\phi_\tau < 1$. It should be noted that f.s.k. systems, for example, need not exhibit this sensitivity

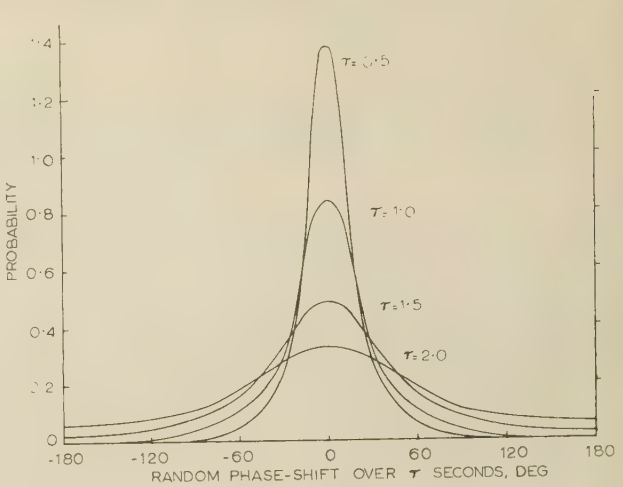


Fig. 1.—Phase stability of the 5 Mc/s WWV carrier received over a short E-layer path, as calculated from the experimental data of R. Price.

to phase perturbations. While coherent or phase-sensitive f.s.k. systems can indeed be constructed, it is easy to show that f.s.k. detectors should usually operate on an energy- or envelope-comparison basis in the presence of fading and noise. It follows from this that phase perturbations which degrade a p.s.k. system seriously will have a smaller, usually negligible, effect on an f.s.k. system.

Eqn. (3) also shows that phase perturbations induced by fading cannot be minimized by radiating more power; they are a function only of ϕ_τ. If the radio path and transmitting frequency are fixed, ϕ_τ is determined by Nature rather than by man. The use of very-short-pulse transmissions ostensibly offers a way to maximize ϕ_τ and thus to 'peak' the p.d.f. of eqn. (3). This method is useful, however, only when the fading is essentially flat or frequency-non-selective. Frequency-selective fading is a manifestation of 'coarse' multi-path propagation, and it is clear that contiguous short-pulse transmissions will be seriously degraded if the relative path delays are a significant fraction of the pulse length.

Finally, readers should note that the foregoing equations and comments are strictly applicable only to fading carrier signals which contain no specular components, and thus exhibit Rayleigh fading. While most of the following Sections deal only with Rayleigh fading, the effect of a specular component will be partially evaluated in Section 8. In the analysis of p.s.k. detection which follows, ϕ_τ is taken as a constant of the environment, and selective fading is ignored (i.e. the data pulses are presumed to be long compared with multi-path time delays). Furthermore, the 'speed' of the fading is constrained to be markedly slower than the data rate; this means that the power spectrum associated with a fading, non-keyed carrier must be markedly narrower than the spectrum of the same carrier when keying is imposed upon it. This constraint has the effect of restricting ϕ_τ to values in the range between, say, 0·9 and 1·0. The restriction to 'slow' fading not only simplifies the analysis, but also is a necessary condition if phase keying is to be feasible in practical situations.

(4) THE BASIC BINARY P.S.K. DETECTOR

The first problem to be attacked is the specification of the basic binary p.s.k. detector. By what processes should transmitted phase shifts be inferred from the received signals? Certain assumptions simplify the answer to this question.

First, additive white noise and fading are assumed to be the

only forms of signal degradation. Section 9 contains comments on the sufficiency of this description of the radio channel. Second, the communication system is assumed to be time-synchronous. In practice, the additive noise should be minimized in an optimum complex-conjugate filter, although this part of the detection process will not be discussed in the paper. The combination of time-synchronism and optimum filtering, when coupled with the restriction to non-selective 'slow' fading, allows each data pulse to be described quite simply, after filtering, in terms of the magnitude and phase of a single vector sample. Third, the two binary states denoted by M (mark) and S (space) are assumed to be equally probable; also, the losses and gains involved in choosing between M and S are assumed to be symmetrical.[6]

Since the receiver is to employ differentially coherent detection, the binary detection process consists in determining whether the current signal pulse connotes a 0° or a 180° phase change with respect to the preceding (stored) signal pulse.

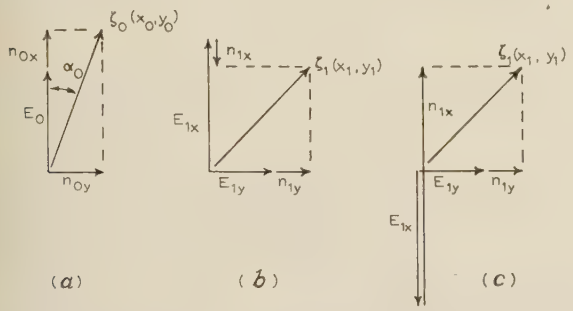


Fig. 2.—Binary phase-keying in a fading, noisy channel.

(*a*) The reference vector, i.e. the preceding signal pulse.
(*b*) The signal vector on the hypothesis of an M (0° shift) transmission.
(*c*) The signal vector on the hypothesis of an S (180° shift) transmission.

Fig. 2 illustrates the situation for a given pair of vectors, ζ_0 and ζ_1. The various vector quantities in the figure are defined as follows:

ζ_0 = Reference vector, which represents the preceding corrupted signal pulse.
ζ_1 = Signal vector, which represents the present corrupted signal pulse.
E_0 = 'Pure' signal component of the reference vector; this is used to orient the axes in Fig. 2.
E_{1x} = 'Pure' signal component of ζ_1 which is in-phase or anti-phase with E_0, depending upon whether M or S was transmitted.
E_{1y} = 'Pure' signal component of ζ_1 which is in quadrature with E_0 due to propagation turbulence.
$n_{0x}, n_{0y}, n_{1x}, n_{1y}$ = Independent, additive, zero-mean, normally-distributed noise vectors of mean power σ_n^2.

If M and S are viewed as two hypotheses, the detection process can be specified via the ratio of their likelihoods.[6, 12] Denoting likelihood functions by L():

$$\frac{\mathrm{L}(M|\zeta_0, \zeta_1)}{\mathrm{L}(S|\zeta_0, \zeta_1)} = \frac{\mathrm{p}(E_0)\mathrm{p}(x_0|E_0)\mathrm{p}(y_0)\mathrm{p}(x_1|E_0, M)\mathrm{p}(y_1)}{\mathrm{p}(E_0)\mathrm{p}(x_0|E_0)\mathrm{p}(y_0)\mathrm{p}(x_1|E_0, S)\mathrm{p}(y_1)}$$

$$= \frac{\mathrm{p}(x_1|E_0, M)}{\mathrm{p}(x_1|E_0, S)} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

If $P_{M/S}$ (preference for M as opposed to S) is a logarithmic measure of the likelihood ratio, then substitution of eqn. (2), after modification to include additive noise effects, yields

$$P_{M/S} \propto x_1 E_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

The process implied by eqn. (5) cannot be implemented directly, since neither E_0 nor x_1 is known. E_0 must be inferred from ζ_0 and x_1 from ζ_1. To see what the inference processes entail, consider the log-likelihood function for E_0 and α_0, where α_0 is the random phase-shift induced in the reference vector by the additive noise (see Fig. 2):

$$\log \mathrm{L}(E_0, \alpha_0|\zeta_0) = \log [\mathrm{p}(x_0|E_0)\mathrm{p}(y_0)]$$

$$\propto \frac{-1}{2\sigma_n^2}(E_0^2 - 2E_0\zeta_0 \cos\alpha_0 + \zeta_0^2) \quad (6)$$

Denoting maximum-likelihood estimators[12] by circumflexes:

$$\frac{\partial}{\partial\alpha_0} \log \mathrm{L}(E_0, \alpha_0|\zeta_0) = -\frac{E_0\zeta_0}{\sigma_n^2} \sin\alpha_0$$

Therefore $\hat{\alpha}_0 = 0$, $E_0\zeta_0 \neq 0$

$$\frac{\partial}{\partial E_0} \log \mathrm{L}(E_0|\zeta_0, \hat{\alpha}_0) = \frac{-1}{\sigma_n^2}(E_0 - \zeta_0)$$

Therefore $\hat{E}_0 = \zeta_0$, where $\hat{E}_0 = \hat{E}_0(\hat{\alpha}_0)$ (7)

Thus, ζ_0 is the best (in the maximum-likelihood sense) estimate of E_0. Similarly, it can be shown that the best estimate of x_1 is proportional to ζ_1. Substituting these estimates in eqn. (5) gives

$$P_{M/S} \propto \zeta_0\zeta_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

Eqn. (8) shows that the decision between M and S should be governed by the polarity of the scalar product of the reference and signal vectors. If $\zeta_0\zeta_1 > 0$, M is chosen; if $\zeta_0\zeta_1 < 0$, S is chosen. While this is certainly not a surprising result, it does justify the use of product detection in p.s.k. demodulation. The transistorized p.s.k. equipment used in the 1957 field test employed synchronous differentially-coherent demodulation via product detection; it is described elsewhere.[13]

The form of the basic binary p.s.k. detector having been specified, the next step is the derivation of the p.d.f. of the product detector output. As shown in Fig. 3, the output, r, is

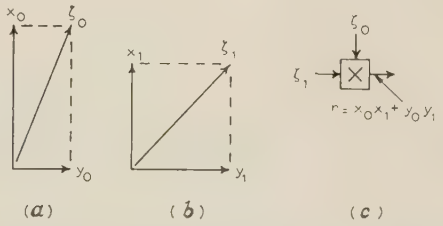


Fig. 3.—Binary phase-keying in a fading, noisy channel.

(*a*) The reference vector.
(*b*) The signal vector.
(*c*) Symbolic representation of the product detector.

the scalar sum of the products of the respective quadrature components. The derivation of the p.d.f. will only be outlined here, for it is tedious in execution although simple in principle.

A transform pair is defined as follows:

$$\mathrm{p}(r|E_0) = \int_{-\infty}^{\infty} \mathrm{P}_r(f|E_0) \exp(2\pi jfr)df \quad . \quad . \quad . \quad (9)$$

$$\mathrm{P}_r(f|E_0) = \int_{-\infty}^{\infty} \mathrm{p}(r|E_0) \exp(-2\pi jfr)dr \quad . \quad . \quad (10)$$

where $\mathrm{p}(r|E_0)$ is the p.d.f. of r, given E_0, and $\mathrm{P}_r(f|E_0)$ is the characteristic function of $\mathrm{p}(r|E_0)$.

Because r is the sum of two independent, random variables,

$$P_r(f|E_0) = P_{x0x1}(f|E_0)P_{y0y1}(f) \quad . \quad . \quad . \quad (11)$$

where, for example,

$$P_{x0x1}(f|E_0) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} p(x_0, x_1|E_0) \exp(-2\pi j f x_0 x_1) dx_1 dx_0 \quad . \quad . \quad . \quad . \quad (12)$$

If $P_{x0x1}(f|E_0)$ and $P_{y0y1}(f|E_0)$ are evaluated by substituting the appropriate normal p.d.f.'s, the following expression for eqn. (10) is obtained:

$$P_r(f|E_0, M) = \left[\frac{1}{1+(2\pi f c\sigma_n)^2}\right] \exp\left\{-2E_0^2\left[\frac{(\pi f)^2(c^2+\phi^2\sigma_n^2)+\pi j f\phi}{1+(2\pi f c\sigma_n)^2}\right]\right\} \quad . \quad . \quad . \quad . \quad (13)$$

where $c^2 = \sigma_n^2 + \sigma_s^2(1-\phi^2)$

σ_s^2 = Mean signal power of either quadrature signal component.

ϕ = Correlation coefficient of either quadrature signal component over the pulse length. (This is ϕ_τ of Section 3.)

Inverse transformation of eqn. (13) and averaging over all values of E_0, which is Rayleigh-distributed, yield

$$p(r|M) = \frac{1}{2(\sigma_s^2+\sigma_n^2)} \exp\left[\frac{r}{\sigma_s^2(1-\phi)+\sigma_n^2}\right], \quad r<0 \quad (14)$$

$$= \frac{1}{2(\sigma_s^2+\sigma_n^2)} \exp\left[\frac{-r}{\sigma_s^2(1+\phi)+\sigma_n^2}\right], \quad r>0 \quad (15)$$

where $p(r|M)$ is the p.d.f. of r, given an M transmission.

The expressions for $p(r|S)$ are identical except for the algebraic sign preceding ϕ in the exponents.

(5) DIVERSITY COMBINING

Since the paper is concerned with phase-shift keying for long-range radiocommunication, it is appropriate at this point to consider diversity reception. In particular, how should the signals in an mth-order diversity receiver be processed? This question can be answered by again examining a likelihood ratio. Suppose that m basic binary p.s.k. detectors operate upon signals which fade independently and have independent additive noise. The kth product detector output is denoted by $r_k^{(+)}$ if it is positive, and by $r_k^{(-)}$ if it is negative. Suppose further that some number, l, of the outputs are positive at a given time, and the remaining $m-l$ are negative. The likelihood ratio for this situation is

$$\frac{L(M|r_1 \ldots r_m)}{L(S|r_1 \ldots r_m)} = \frac{\prod_{k=1}^{l} p(r_k^{(+)}|M) \prod_{k=l+1}^{m} p(r_k^{(-)}|M)}{\prod_{k=1}^{l} p(r_k^{(+)}|S) \prod_{k=l+1}^{m} p(r_k^{(-)}|S)} \quad (16)$$

If a logarithmic measure of the likelihood ratio is denoted by $P_{M/S}^{(m)}$ and the appropriate p.d.f.'s are substituted, it follows that

$$P_{M/S}^{(m)} \propto \sum_{k=1}^{l} r_k^{(+)} - \sum_{k=l+1}^{m} |r_k^{(-)}| = \sum_{k=1}^{m} r_k \quad . \quad . \quad (17)$$

Thus, the combination of the product detector outputs should be achieved by linear addition, and decisions should be determined by positive and negative excursions of the sum.

It is interesting to note that this combining method agrees with the results of Pierce, who investigated ideal coherent detection of fading f.s.k. signals.[14] Pierce found that when the phases and amplitudes of f.s.k. signals can be predicted before detection, the samples from each receiver should be weighted by the expected amplitude, and the weighted samples should be added. The weighting operation prescribed by Pierce is implicit in differentially-coherent product detection.

(6) ERROR PROBABILITIES IN DIVERSITY RECEPTION OF BINARY P.S.K. SIGNALS

Error probabilities can now be derived for mth-order diversity reception of binary p.s.k. signals. Because of symmetry in the decision rule and signal statistics, only the p.d.f.'s for M (or S, but not both) transmissions are required to derive the various expressions in this and following Sections.

Eqn. (13) gives the characteristic function of the p.d.f. of the output of one product detector. Since m product detectors operate on m independently fading paths, one may write by inspection

$$P_s(f|M, W_m) = [1+(2\pi f c\sigma_n)^2]^{-m} \exp\left\{-2W_m\left[\frac{(\pi f)^2(c^2+\phi^2\sigma_n^2)+\pi j f\phi}{1+(2\pi f c\sigma_n)^2}\right]\right\} \quad . \quad . \quad . \quad . \quad (18)$$

where $P_s(f|M, W_m)$ = Characteristic function of $p(s|M, W_m)$.

s = Combined output of the m product detectors.

$$W_m = \sum_{k=1}^{m} E_{0k}^2.$$

Expressions for $p(s|M, m)$, where m is the order of diversity, can be found by inverse transformation of eqn. (18) and by averaging over all values of W_m. Because the various E_0 terms fade independently, the p.d.f. of W_m (which is required for the averaging) is[14]

$$p(W_m) = \left(\frac{1}{2\sigma_s^2}\right)^m \frac{(W_m)^{m-1}}{(m-1)!} \exp\left(-\frac{W_m}{2\sigma_s^2}\right), \quad 0 \leqslant W_m, \; m = 1, 2, 3, \ldots \quad (19)$$

Thus, for example,

$$p(s|M, 2) = \left[\frac{1}{2(\sigma_s^2+\sigma_n^2}\right]^2 \exp\left[\frac{s}{\sigma_s^2(1-\phi)+\sigma_n^2}\right] \left[\frac{\sigma_n^4+2\sigma_n^2\sigma_s^2+\sigma_s^4(1-\phi^2)}{\sigma_s^2+\sigma_n^2} - s\right], \; s<0 \quad (20)$$

Average error probabilities are found by integrating $p(s|M, m)$ from $-\infty$ to 0. For the non-diversity and dual diversity cases, the following results obtain:

$$P_{\epsilon,1} = \frac{\rho(1-\phi)+1}{2(\rho+1)} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (21)$$

$$P_{\epsilon,2} = (P_{\epsilon,1})\left[\frac{\rho^2(2-\phi-\phi^2)+\rho(4-\phi)+2}{2(\rho+1)^2}\right] \quad (22)$$

where $P_{\epsilon,m}$ = Probability of an error in mth-order diversity reception of binary p.s.k. signals.

$\rho = \sigma_s^2/\sigma_n^2$, a mean signal/noise ratio.

The structure of the error probabilities shows that errors are induced by two sources—noise and propagation. When $\phi \simeq 1$ and ρ is small, the error probabilities decrease with increasing signal/noise ratio, as one might expect. Unlimited increases in ρ are useless, however, because at some point errors

nduced by the transmission medium begin to exceed those caused by additive noise. The irreducible error probabilities characteristic of binary p.s.k. transmission in Rayleigh-fading media, which were predicted in Section 3, may be found by taking the limits of eqns. (21) and (22), as ρ approaches infinity.

$$P_{\epsilon,1}|_{\rho=\infty} = \frac{1-\phi}{2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (23)$$

$$P_{\epsilon,2}|_{\rho=\infty} = \left(\frac{1-\phi}{2}\right)\left[1 - \frac{\phi(1+\phi)}{2}\right] \quad . \quad . \quad (24)$$

Diversity reception is obviously useful in reducing errors caused by both noise and propagation. Although the analysis has been restricted to paths which fade independently, Pierce's work on correlated fading implies that the diversity gains will still be substantial when moderate correlation exists between the paths.[14]

(7) ASYMPTOTIC ERROR PROBABILITIES IN DIVERSITY RECEPTION OF QUATERNARY P.S.K. SIGNALS

The investigation has dealt thus far with noisy binary p.s.k. signals. The analysis of quaternary p.s.k. transmission in this Section is limited to noiseless signals; the error probabilities which are derived are thus attributable solely to the fading medium. While solutions including noise effects can probably be found, the asymptotic values given here are easy to obtain and are of considerable practical interest.

The basic receiver is again assumed to be a time-synchronous differentially-coherent device. The dual product detectors required for quaternary p.s.k. detection are shown symbolically in Fig. 4, together with vector diagrams of the signals. Diversity

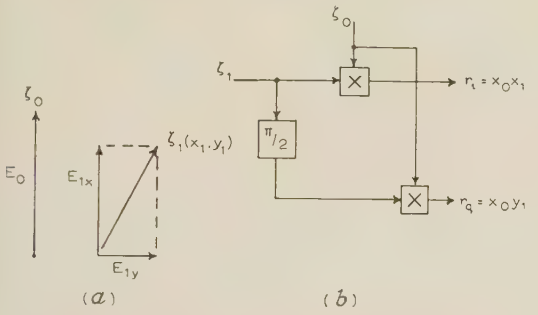


Fig. 4.—Quaternary phase-keying in a fading, noiseless channel.
(*a*) The reference and signal vectors.
(*b*) Symbolic representation of the quaternary p.s.k. detector.

combining is effected by linear addition of the product detector outputs. Because two bits of information are conveyed in quaternary p.s.k. signals by transmitted phase shifts of 0°, 180° or ±90°, the following rules govern the decision process when the four phase states are equally probable:

$$(|s_i| > |s_q|, \quad s_i > 0) \rightarrow 0°$$
$$(|s_i| > |s_q|, \quad s_i < 0) \rightarrow 180°$$
$$(|s_q| > |s_i|, \quad s_q > 0) \rightarrow 90°$$
$$(|s_q| > |s_i|, \quad s_q < 0) \rightarrow -90°$$

where the symbol $\rightarrow$ means 'implies'.

s_i = Combined output of the i (in-phase) product detectors.
s_q = Combined output of the q (quadrature) product detectors.

In contrast to the binary case, errors need not result in a total loss of information; therefore, expressions must be found for both 1-bit and 2-bit errors. These expressions follow immediately from the foregoing decision rules:

$$P_{\epsilon 0,m} = 2\int_0^\infty p(W_m)\int_0^\infty p(s_i|W_m)\int_0^{s_i} p(s_q|W_m)ds_q ds_i dW_m \cdot \quad (25)$$

$$P_{\epsilon 2,m} = 2\int_0^\infty p(W_m)\int_{-\infty}^0 p(s_i|W_m)\int_0^{-s_i} p(s_q|W_m)ds_q ds_i dW_m \cdot \quad (26)$$

$$P_{\epsilon 1,m} = 1 - (P_{\epsilon 0,m} + P_{\epsilon 2,m}) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (27)$$

where $P_{\epsilon 0,m}$ = Probability of a 'zero error', i.e. a correct decision, in mth-order diversity reception of quaternary p.s.k. signals.
$P_{\epsilon 2,m}$ = Probability of a 2-bit error.
$P_{\epsilon 1,m}$ = Probability of a 1-bit error.

The various p.d.f.'s in eqns. (25) and (26) are either given earlier [see eqn. (19)] or can be found readily by the characteristic-function method outlined in Section 4. Eqns. (25) and (26) can be evaluated for any reasonable order of diversity by successive integrations by parts and by use of the identity

$$\int_0^\infty \exp\left(-\frac{c^2r^2}{2}\right)\int_0^{kr}\exp\left(-\frac{t^2}{2}\right)dtdr = \frac{1}{c}\arctan\left(\frac{k}{c}\right)$$

For the non-diversity and dual-diversity cases, the following expressions obtain for the noiseless but fading channel:

$$P_{\epsilon 1,1} = \frac{1}{2} - 2\phi(2-\phi^2)^{-1/2}\frac{1}{\pi}\arctan\left[\phi(2-\phi^2)^{-1/2}\right] \quad . \quad (28)$$

$$P_{\epsilon 2,1} = \frac{1}{4} - \phi(2-\phi^2)^{-1/2}\left\{\frac{1}{2} - \frac{1}{\pi}\arctan\left[\phi(2-\phi^2)^{-1/2}\right]\right\} \quad (29)$$

$$P_{\epsilon 1,2} = P_{\epsilon 1,1} - \frac{2\phi(1-\phi^2)}{2-\phi^2}\left\{\frac{\phi}{2\pi} + \frac{1}{\pi}(2-\phi^2)^{-1/2}\arctan\left[\phi(2-\phi^2)^{-1/2}\right]\right\} \quad (30)$$

$$P_{\epsilon 2,2} = P_{\epsilon 2,1} - \frac{\phi(1-\phi^2)}{2-\phi^2}\left((2-\phi^2)^{-1/2}\left\{\frac{1}{2} - \frac{1}{\pi}\arctan\left[\phi(2-\phi^2)^{-1/2}\right]\right\} - \frac{\phi}{2\pi}\right) \quad (31)$$

Thus, propagation-induced errors occur in quaternary p.s.k. signals whenever $\phi < 1$, even though the signal/noise ratio is infinite. As in the binary case, diversity is useful in minimizing these errors.

(8) THE EFFECT OF A SPECULAR COMPONENT

All the preceding results have been predicted upon Rayleigh fading. This implies that the received signal contains no steady or specular component whose phase is constant (except for transmitted shifts and additive noise effects) compared to a stable local reference oscillator. It is known from experience, however, that fading signals frequently do not exhibit ideal Rayleigh statistics, and sometimes follow a Rice distribution.[5, 7, 10] This situation will be investigated for non-diversity binary p.s.k. signals in a noiseless channel.

Fig. 5 shows a specular component, E_s, as well as the fluctuating components E_x and E_y used previously; note that the reference axes are now fixed by E_s. By slightly modifying the

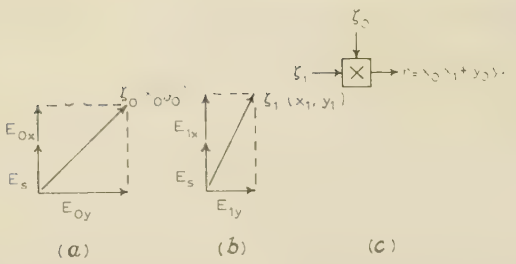


Fig. 5.—Binary phase-keying in a noiseless channel with fading superposed on a specular component.

(*a*) The reference vector.
(*b*) The signal vector.
(*c*) The product detector.

characteristic-function method outlined earlier, the following p.d.f. can be derived for the output of the product detector:

$$p(r|M) = \frac{1}{2\sigma_s^2} \exp\left[\frac{-E_s^2(1-\phi) + 2r}{2\sigma_s^2(1-\phi)}\right], \quad r < 0 \qquad (32)$$

The average error probability is found by integrating eqn. (32) from $-\infty$ to 0, giving

$$P_\epsilon = \left(\frac{1-\phi}{2}\right) \exp\left(-\frac{E_s^2}{2\sigma_s^2}\right) \quad . \quad . \quad . \quad (33)$$

The exponent of eqn. (33) might be termed a 'propagation' signal/noise ratio, for it involves the power in the specular component and the mean fluctuation power. When $E_s = 0$, eqn. (33) reduces to eqn. (23). When the fluctuation power is zero, the error probability vanishes in the noiseless case. Eqn. (33) indicates that a strong specular component can effect a marked reduction in the errors due to propagation turbulence. Unfortunately, values of E_s/σ_s significantly greater than unity occur but infrequently on most long-range radio circuits.

It is worth noting that the model shown in Fig. 5 applies to the case of a non-fading signal in additive noise if $\phi = 0$ and σ_s^2 is interpreted as σ_n^2, the mean noise power. With these modifications, eqn. (33) reduces to

$$P_\epsilon|_{non\text{-}fading} = \tfrac{1}{2} \exp\left(-\frac{E_s^2}{2\sigma_n^2}\right) \quad . \quad . \quad . \quad (34)$$

This result was derived by Lawton in his study of binary p.s.k. and other digital modulation techniques under non-fading and additive noise conditions.[15]

(9) FIELD-TEST RESULTS

The 1957 field test mentioned in the Introduction was conducted over a 5000-mile high-frequency radio path from the Hawaiian Islands to Fort Monmouth, New Jersey, U.S.A. The equipments involved were a 16-channel f.s.k. terminal and a 12-channel quaternary p.s.k. terminal. The f.s.k. system was of the asynchronous narrow-band limiter-discriminator type with 'ratio squared' diversity combining, while the p.s.k. system employed synchronous differentially-coherent detection with additive diversity combining. Single-sideband radio equipment and dual space-diversity reception were used throughout the test. About $1{\cdot}4 \times 10^7$ words of teleprinter copy were received, and the error statistics were subsequently analysed on a digital computer. Plots of many of the results and a comprehensive description of the test are published elsewhere.[1]

A striking result which emerged from the field test was the comparative performance of the two systems as a function of signal/noise ratio. As noted earlier, the p.s.k. system demonstrated a fairly constant superiority at relatively low ratios (less than about 30 dB, measured in a 3 kc/s bandwidth containin 12 channels), while the f.s.k. system showed a superiority a high ratios which increased steadily with the ratio. Th results of the preceding Sections, coupled with existin theoretical analyses of f.s.k. systems, e.g. Reference 14, seen to explain this phenomenon as follows. At low signal/nois ratios, noise was apparently the main cause of errors an the smaller noise bandwidth of the p.s.k. system gave i a relative advantage. At high signal/noise ratios, however, th p.s.k. system apparently had reached the error-rate 'floor' se by the phase stability of the medium; thus, only the f.s.k. systen profited from further increases in signal power. Theoretica curves were fitted to the comparative-performance data to obtai crude estimates of the ϕ parameter.

The comparative results discussed above were presented in th form of variations of the error-rate ratio of the two systems wit signal/noise ratio. (It should be noted that this type of presenta tion requires very careful data processing if the results are to li within a reasonable confidence interval.) Other results were th individual performances of the two systems. These were les reassuring from a theoretical viewpoint, for the experimenta data did not fit theoretical curves of error-rate versus signal/nois ratio.

Two major discrepancies were noted. First, the slopes of th experimental curves were considerably smaller than expected about 20–25 dB of signal power was required, on the average to reduce the error rates by a factor of ten. Second, both systems seemed to approach an irreducible error rate at very high signal/noise ratios, though this asymptotic behaviour wa more marked in the p.s.k. system. This author believes tha these discrepancies between theory and experiment in the indi vidual performances are directly attributable to the assumption underlying the theoretical analyses. In particular, the descrip tion of additive disturbances in the long-range radio environ ment in terms of stationary, normally distributed white noise seems unrealistic. The assumption of white noise has, however, two advantages: it is mathematically tractable, and one can be assured that white noise is always present, although it may not be the dominant form of additive disturbance.

Why did theoretical analyses based on a seemingly unrealistic assumption succeed in predicting the comparative performance of the two systems, and yet fail to predict the individual performance of either system? The author suggests that the answer lies in the nature of the systems. The characteristic of phase-shift keying which makes it attractive is bandwidth conservation; thus, one might expect a p.s.k. system to combat most forms of wide-band additive disturbance more efficiently than an f.s.k. system. If one views impulse noise and many other types of transient phenomena simply as wide-band additive disturbances, it follows that analyses based on white noise plus fading should predict f.s.k.–p.s.k. comparative performance quite accurately, although the same analyses may be useless for predicting the individual performance of either system.

(10) AVERAGED-PHASE REFERENCES

It was noted in Section 2 that many methods have been proposed for transmitting phase references; therefore the restriction to differentially-coherent detection (phase comparison with the preceding pulse) which obtains throughout the theoretical analysis may seem arbitrary. One can envisage, for example, p.s.k. systems which ostensibly derive 'clean' phase references by averaging several of the preceding signal pulses. The statistical procedures used previously can be extended to analyse phase-averaging schemes, and, in fact, an optimum 'memory' function can be found as a function of $\phi(\tau)$ and the signal/noise ratio

which minimizes the error probability. As one might expect, the mathematics shows that a large number of data pulses should be used to establish a phase reference when additive noise is the preponderant source of errors, but that only the preceding pulse should be used when errors arise mainly from propagation fluctuations.

Although a perfect phase reference can never be established for the detection of signals transmitted through a fading medium, it is worth while to determine the maximum improvement which a perfect reference would afford. It is well known that the error probability for ideal coherent detection of a non-fading binary-phase-keyed signal in white noise is

$$(P)_{ideal} = \tfrac{1}{2} - \int_0^{E/\sigma_n} \frac{1}{\sqrt{2\pi}} \exp\left(-\frac{t^2}{2}\right) dt$$

Since fading will affect only the signal amplitude in the ideal-reference case, the error probability for ideal non-diversity reception of a Rayleigh-fading signal is

$$(P_f)_{ideal} = \tfrac{1}{2} - \int_0^{\infty} \frac{E}{\sigma_s^2} \exp\left(-\frac{E^2}{2\sigma_s^2}\right) \int_0^{E/\sigma_n} \frac{1}{\sqrt{2\pi}} \exp\left(-\frac{t^2}{2}\right) dt dE$$

$$= \tfrac{1}{2}\left[1 - \left(\frac{\rho}{\rho+1}\right)^{1/2}\right] \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (35)$$

Comparison of eqn. (35) with eqn. (21) under the necessary condition that $\rho(1-\phi) \ll 1$ shows that a differentially-coherent p.s.k. system requires about 3 dB more signal power than a p.s.k. system using an ideal phase reference in order to equal the performance of the latter. Thus any physically realizable scheme for deriving a 'clean' phase reference can approach a 3 dB improvement only under the most favourable conditions. Furthermore, it is not difficult to show that a phase-averaging scheme which has too long a 'memory' (compared to the phase stability of the medium) can degrade the performance of a p.s.k. system severely.

These comments on phase-averaging strictly apply only to Rayleigh-fading signals. If a strong specular component is present, phase-averaging is obviously more feasible. Lawton has shown, however, that in the limiting case of no fading the advantage offered by ideal coherent detection is only in the order of 1 dB.[15]

(11) SEQUENTIAL ERRORS

It is important to understand that the preceding analysis deals only with independent errors. The error probabilities derived earlier can be applied in two contexts. In the first, consider a multiplicity of p.s.k. receiving systems (where a system may be an mth-order diversity receiver) processing information from statistically identical but independently variable paths. The error probabilities are then a property of all sets composed of one decision selected at random from each system. In the second context, consider a single receiving system emitting sequential decisions spaced τ seconds apart. The error probabilities are then a property of all sets composed of decisions spaced $K\tau$ seconds apart, where $\phi(K\tau) \simeq 0$. These restrictions are necessary because error-clustering has been neglected. That errors will indeed tend to occur in groups or clusters is intuitively evident from the high degree of coherence required for acceptable phase-shift keying. The transition probabilities which describe error clusters are very important in coding studies, but their mathematics are less tractable in the p.s.k. case than for the f.s.k. case. The error probabilities derived in the paper are applicable to the 1957 field test, wherein character (non-redundant multi-digit group) error statistics were collected, only on the rather questionable assumption that the correlation coefficient over the character length was small.

(12) COMMENTS

The preceding analysis shows theoretically that p.s.k. transmission through fading channels is subject to unavoidable degradation, and that this degradation cannot be overcome by radiating more power. Many questions remain unanswered, however. The analysis tells nothing about frequency-selective propagation effects or sequential error statistics. One might even question whether the fluctuations in received phase-keyed signals can be described by stationary, ergodic statistical models; perhaps the underlying physical processes are non-stationary even over short time intervals. Nevertheless, from a practical viewpoint, the limited experimental data available imply that p.s.k. systems are indeed limited by the phase stability of radio channels.

What can be done to optimize long-range radio systems using phase-shift keying? From the comments in Section 3 on short-pulse transmissions, it is clear that p.s.k. systems should use the shortest pulse length possible, subject to the limitations imposed by 'coarse' multi-path time delays. Thus, the signalling speed should be a compromise which minimizes the sum of the errors caused by phase fluctuations and multi-path delays. This means that the signalling speed should be adjustable to cope with a variety of propagation conditions, and this in turn requires terminal equipment which is not only flexible but which can also monitor continuously the characteristics of the radio path. The design of channel-matching radio systems (i.e. systems which adjust themselves to propagation conditions) is in its infancy; one such system for highly redundant signals is described in Reference 16.

However, there is reason to believe that redundant coding may be more effective than channel matching for improving p.s.k. performance, although the development of simple codes for phase-shift keying poses some unusual problems. For example, the very nature of the differentially-coherent detection process promotes error doublets. This characteristic means that simple codes, such as single parity checks and constant-ratio groups, must be interlaced if they are to be effective. Moreover, it was noted in Section 11 that p.s.k. errors are bound to occur in clusters, because propagation fluctuations must occur slowly, compared with the signalling speed, for phase-shift keying to be feasible; this further complicates the design of simple codes, although it is advantageous from the viewpoint of pure communication theory.

It is quite possible that phase-averaging techniques may be unexpectedly useful in both channel-matching and redundantly coded p.s.k. systems. This follows because phase-averaging over several data pulses can provide not only a measure of the channel stability, but also a means of controlling error-clustering to some extent. Thus it might actually be advantageous to use an averaged-phase reference which (incidentally) increases the probability of independent errors in order to enhance the efficiency of a particular redundant code.

In summary, the author believes that p.s.k. systems offer at present a means of conserving bandwidth, but that they do not necessarily offer greater reliability for long-range radio service than more conventional systems. Because the requirements for both accuracy and capacity in long-range data systems are likely to increase markedly in the foreseeable future, efforts should be directed toward the improvement of p.s.k. systems. In the author's opinion, worth-while improvements can be effected by

adjusting signalling speeds and detection mechanisms so as to match the channel, and by efficient redundant coding. The two types of technique are related, and both require more knowledge about the statistics of actual radio channels than is now available. However, some of the results reported here—eqns. (23) and (33), for example—suggest that certain relevant propagation characteristics can be measured quite easily with a simple p.s.k. receiver and an unmodulated, stable, high-power transmitter.

(13) ACKNOWLEDGMENTS

Much of the work reported here was done while the author was a Project Officer at the U.S. Army Signal Research and Development Laboratory, Fort Monmouth, New Jersey, U.S.A. The author particularly acknowledges the encouragement afforded by Messrs. R. A. Kulinyi, B. Goldberg, and H. E. White of that Laboratory, and thanks the U.S. Department of Defense for permission to publish the work. Professors E. C. Cherry and G. A. Barnard of the Imperial College of Science and Technology provided valuable insight into some of the statistical concepts used in the presentation.

(14) REFERENCES

(1) Brennan, A. T., Goldberg, B., and Eckstein, A.: 'Comparison of Multi-Channel Radioteletype Systems Over a 5000-Mile Ionospheric Path', *National Convention Record of the Institute of Radio Engineers*, 1958, **8**, p. 254.

(2) Costas, J. P.: 'Phase-Shift Radio Teletype', *Proceedings of the Institute of Radio Engineers*, 1957, **45**, p. 16.

(3) Ratcliffe, J. A.: 'Diffraction from the Ionosphere and the Fading of Radio Waves', *Nature*, 1948, **162**, p. 4105.

(4) Booker, H. G., Ratcliffe, J. A., and Shinn, D. H.: 'Diffraction from an Irregular Screen, with Applications to Ionospheric Problems', *Philosophical Transactions of the Royal Society* (*A*), 1950, **262**, p. 579.

(5) McNicol, R. W. E.: 'The Fading of Radio Waves of Medium and High Frequencies', *Proceedings I.E.E.* Paper No. 891 R, November, 1949 (**96**, Part III, p. 517)

(6) Woodward, P. M.: 'Probability and Information Theory with Applications to Radar' (Pergamon Press, 1953).

(7) Rice, S. O.: 'Mathematical Analysis of Random Noise' *Bell System Technical Journal*, 1945, **24**, p. 45.

(8) Bramley, E. N.: 'Diversity Effects in Spaced-Aerial Reception of Ionospheric Waves', *Proceedings I.E.E.*, Paper No. 1062 R, January, 1951 (**98**, Part III, p. 19).

(9) Price, R.: 'The Autocorrelogram of a Complete Carrier Wave Received over the Ionosphere at Oblique Incidence', *Proceedings of the Institute of Radio Engineers*, 1957, **45**, p. 879.

(10) Brennan, D. G., and Phillips, M. L.: 'Phase and Amplitude Variability in Medium-Frequency Ionospheric Transmissions', Technical Report No. 93, The Lincoln Laboratory, Lexington, Massachusetts, U.S.A., 1957.

(11) Hagfors, T., and Landmark, B.: 'Simultaneous Variation of Amplitude and Phase of Gaussian Noise, with Applications to Ionospheric Forward Scatter Signals', *Proceedings I.E.E.*, Paper No. 2743 R, November, 1958 (**105** B, p. 555).

(12) Fisher, R. A.: 'Statistical Methods and Scientific Inference' (Oliver and Boyd, 1956).

(13) Doelz, M. L., Heald, E. T., and Martin, D. L.: 'Binary Data Transmission Techniques for Linear Systems', *Proceedings of the Institute of Radio Engineers*, 1957, **45**, p. 656.

(14) Pierce, J. N.: 'Theoretical Diversity Improvement in Frequency-Shift Keying', *ibid.*, 1958, **46**, p. 903.

(15) Lawton, J. G.: 'Comparison of Binary Data Transmission Systems', *Proceedings of The Second National Conference on Military Electronics*, 1958, p. 54; see also *Proceedings of the Institute of Radio Engineers*, 1959, **47**, p. 333.

(16) Price, R., and Green, P. E., Jr.: 'A Communication Technique for Multipath Channels', *Proceedings of the Institute of Radio Engineers*, 1958, **46**, p. 555.

621.372.542.29

The Institution of Electrical Engineers
Paper No. 3121 E
Jan. 1960
©

AN ANALYSIS OF A TYPE OF COMB FILTER

By A. G. J. MacFARLANE, B.Sc., Graduate.

(*The paper was first received* 24*th April, and in revised form* 14*th July*, 1959.)

SUMMARY

Several basic arrangements of delay unit, combining unit and feedback unit are shown to have amplitude/frequency response characteristics of the type which has given rise to the name *comb filter*. Such filters are useful for the selection or rejection of a train of uniformly spaced regular pulses. The response of the most suitable configurations to such a train of pulses and to a Gaussian noise input is investigated with particular reference to the improvement in signal/noise ratio possible with this type of filter. The practical realization of such filters is discussed and the most suitable forms of practical comb filter are indicated, together with a brief discussion of their incorporation into a radar system.

LIST OF PRINCIPAL SYMBOLS

$Y(p)$ = Transfer function.
$Y(j\omega)$ = Amplitude/frequency response characteristic.
V_i = Input-signal voltage function, volts.
$\bar{V}_i$ = Transform of input-signal function.
V_o = Output-signal voltage function, volts.
$\bar{V}_o$ = Transform of output-signal function.
β = Feedback loop gain.
N = Number of signal pulses in input-signal train.
T = Delay time of delay unit, sec.
$S_i(t)$ = Input-signal sequence function, volts.
$S_o(t)$ = Output-signal sequence function, volts.
$\bar{S}_i$ = Transform of input-signal sequence.
$\bar{S}_o$ = Transform of output-signal sequence.
I_{SN} = Signal/noise ratio improvement of integrating comb filter.
$\mathscr{L}^{-1}$ = Inverse Laplace transform.

(1) INTRODUCTION

A *comb filter*[1,2] is a special type of filter designed to select or suppress signals having a known periodic frequency structure. The name is derived from the resemblance of the amplitude/frequency response characteristic to the teeth of a comb, a resemblance which may be seen in Figs. 3 and 5.

An obvious method of achieving such a frequency response characteristic is by paralleling a number of single-peaked response filters whose peaks are separated in frequency by the desired amounts.[20] Such a solution is usually impracticable as the number of response peaks required may be several thousand. The input to the filter is a periodic function of time, and an alternative approach to the problem of selective filtering is to delay input signals by multiples of the basic repetition period and subsequently combine the delayed groups of signals together. Two basic methods of achieving this are shown in Fig. 1. In the arrangement of Fig. 1(*b*), a feedback circuit must be added to control the stability of the filter if any active elements are included in the delay unit, as a positive-feedback loop has been formed.

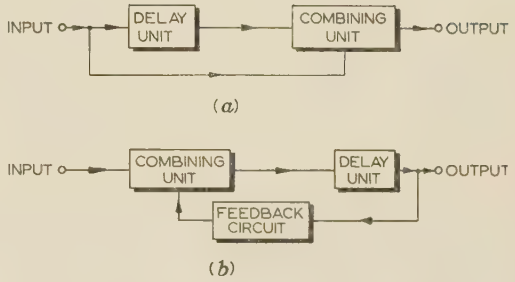


Fig. 1.—Combining schemes for comb filter.

The practical form taken by this type of comb filter is determined by the type of delay or storage element used. The simplest types of comb filter use a propagating type of delay unit, usually a quartz or mercury[11] ultrasonic delay line. Any type of storage may be used, the various possibilities being discussed in Section 4. In analysing the performance of the filter, an important distinction must be made between those types which use signal storage in a propagating medium (or any other type of storage having continuous input and output), and those using a storage element whose information storage capacity will saturate after a certain time when fed with a continuous input, e.g. an electrostatic storage tube.[5, 7, 8, 10] With this second type of storage element, which has a finite information-handling capacity, the input information must be combined in sequences of length determined by the storage capacity. The behaviour of such a type of comb filter is found to differ from that of the simpler type using storage with continuous input and output.

Comb filters have been used in radars[23] having a constant pulse-repetition frequency. Signals at the output of such a radar consist of a series of equally spaced pulses embedded in thermal noise or interference whose level sets the working range of the radar. The comb filter may be designed to select target pulses spaced at the radar pulse-repetition interval, while largely rejecting noise and interference, and its use in this manner is termed *signal integration*, successive pulses being added together or integrated. Alternatively, the comb filter may be designed to reject frequencies which are multiples of the basic radar pulse-repetition frequency, its use in this way enabling moving targets having Doppler-modulated responses to be distinguished from stationary returns.[23, 27] Such a system is termed a *moving target indicator* or *canceller*. The paper is confined to an analysis of the performance of the type of comb filter shown in Fig. 1; as comb filters find their main use in radar, the input signals considered are idealized versions of those occurring in radar. A filter designed to reject a train of equally spaced pulses will be termed a *canceller*, and one designed to select such a train of pulses, an *integrator*.

(2) TRANSFER FUNCTION AND FREQUENCY RESPONSE OF BASIC COMB FILTER ARRANGEMENTS

The basic arrangements of Fig. 1 are considered with additive and subtractive combining units, and the Laplace transform

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. MacFarlane was formerly in the Department of Electrical Engineering, Royal Technical College, Salford, and is now in the Department of Electrical Engineering, Queen Mary College, University of London.

method of analysis is adopted. The transfer function of an ideal delay unit of delay time T seconds can be deduced from the shifting theorem[3] and is

$$Y(p) = \varepsilon^{-pT} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The four possible arrangements are shown diagrammatically in Fig. 2. In this initial analysis, the feedback circuit of Fig. 1(*b*) is taken as having a transfer function which is a pure multiplier, β, a real number less than unity.

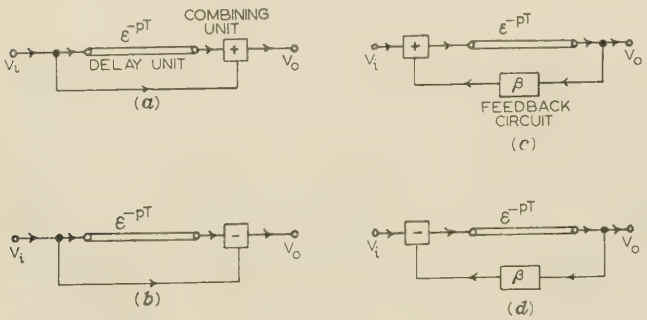


Fig. 2.—Basic comb filter arrangements.
(*a*) Type A filter.
(*b*) Type B filter.
(*c*) Type C filter.
(*d*) Type D filter.

The transfer functions of these four basic arrangements can be deduced as follows: for Fig. 2(*a*),

$$\bar{V}_o = \bar{V}_i + \bar{V}_i\varepsilon^{-pT} \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Thus
$$\bar{V}_o = \bar{V}_i(1 + \varepsilon^{-pT}) \quad . \quad . \quad . \quad . \quad . \quad (3)$$

and
$$Y_a(p) = \frac{\bar{V}_o}{\bar{V}_i} = 1 + \varepsilon^{-pT} \quad . \quad . \quad . \quad . \quad . \quad (4)$$

Similarly, for Fig. 2(*b*),

$$Y_b(p) = 1 - \varepsilon^{-pT} \quad . \quad . \quad . \quad . \quad . \quad (5)$$

For the arrangement of Fig. 2(*c*),

$$(\bar{V}_i + \beta\bar{V}_o)\varepsilon^{-pT} = \bar{V}_o \quad . \quad . \quad . \quad . \quad . \quad (6)$$

Thus
$$\bar{V}_o(1 - \beta\varepsilon^{-pT}) = \bar{V}_i\varepsilon^{-pT} \quad . \quad . \quad . \quad . \quad (7)$$

and
$$Y_c(p) = \frac{\bar{V}_o}{\bar{V}_i} = \frac{\varepsilon^{-pT}}{1 - \beta\varepsilon^{-pT}} \quad . \quad . \quad . \quad . \quad (8)$$

Similarly, for the arrangement of Fig. 2(*d*),

$$Y_d(p) = \frac{\varepsilon^{-pT}}{1 + \beta\varepsilon^{-pT}} \quad . \quad . \quad . \quad . \quad . \quad (9)$$

The amplitude/frequency response characteristics for the four arrangements can now be obtained from the transfer functions by setting $p = j\omega$ and taking the modulus of the resulting complex transfer function.

Proceeding thus, for the arrangement of Fig. 2(*a*) we have

$$Y_a(j\omega) = 1 + \varepsilon^{-j\omega T} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Thus
$$Y_a(j\omega) = 1 + \cos\omega T - j\sin\omega T$$

whence
$$|Y_a(j\omega)| = 2\left|\cos\frac{\omega T}{2}\right| \quad . \quad . \quad . \quad . \quad (11)$$

If M filters of this type are unilaterally cascaded, the overall frequency response characteristic will be

$$|Y_{aM}(j\omega)| = 2^M\left|\cos^M\frac{\omega T}{2}\right| \quad . \quad . \quad . \quad . \quad (12)$$

Similarly, for the arrangement of Fig. 2(*b*),

$$|Y_b(j\omega)| = 2\left|\sin\frac{\omega T}{2}\right| \quad . \quad . \quad . \quad . \quad (13)$$

and for M in a unilaterally cascaded arrangement,

$$|Y_{bM}(j\omega)| = 2^M\left|\sin^M\frac{\omega T}{2}\right| \quad . \quad . \quad . \quad . \quad (14)$$

Considering the transfer function of eqn. (8) for the arrangement of Fig. 2(*c*) we have

$$Y_c(j\omega) = \frac{\varepsilon^{-j\omega T}}{1 - \beta\varepsilon^{-j\omega T}} \quad . \quad . \quad . \quad . \quad (15)$$

and thus, since $|\varepsilon^{-j\omega T}| = 1$,

$$|Y_c(j\omega)| = \left|\frac{1}{(1 - \beta\cos\omega T) + j\beta\sin\omega T}\right|$$

whence
$$|Y_c(j\omega)| = \frac{1}{\sqrt{(1 + \beta^2 - 2\beta\cos\omega T)}} \quad . \quad . \quad (16)$$

and a similar argument for the arrangement of Fig. 2(*d*) gives

$$|Y_d(j\omega)| = \frac{1}{\sqrt{(1 + \beta^2 + 2\beta\cos\omega T)}} \quad . \quad . \quad (17)$$

The functions given by eqns. (16) and (17) have maxima and minima of value $1/(1 - \beta)$ and $1/(1 + \beta)$ respectively, at angular frequencies given by

$$\omega = \frac{k\pi}{T} \quad . \quad . \quad . \quad . \quad . \quad . \quad (18)$$

where k is zero or integral. In eqn. (16), maxima occur at zero or even values of k, and in eqn. (17) at odd values of k. The minima occur at odd values of k in eqn. (16) and at zero and even values of k in eqn. (17).

In Figs. 3(*a*)–(*d*) are shown the amplitude/frequency response characteristics given by eqns. (12), (14), (16) and (17). It is never necessary, as discussed later, to cascade filters arranged as in Figs. 2(*c*) and (*d*), and consequently the effect of cascading these filters has not been considered. For brevity in subsequent discussions of the relative merits of the various arrangements of Figs. 2(*a*)–(*d*), these are subsequently referred to as filters of type A, B, C and D respectively. Inspection of the frequency response characteristics of Fig. 3 shows the characteristic shape which has given rise to the name of comb filter for devices of this type.

(2.1) Group-Handling Type of Filter

Filters arranged as in Figs. 2(*c*) and (*d*) require the delay or storage unit to handle the superimposed result of many signals. All types of storage unit, however, cannot handle the input signals on a continuous basis. In an electrostatic storage tube,[5, 7, 8, 10] for example, information is stored in terms of an electric charge distribution over a finite area. With a continuous input of information, the finite storage capacity is completely filled after a certain time and no further input information can be accepted. It is necessary to erase the contents of the store partially or entirely before fresh input information can be accepted. If this type of storage unit is used, therefore, it is necessary to combine the input information in sequences or groups. The length of the sequence is determined by the information storage capacity of the storage unit, and is arranged to be a multiple of the fundamental repetition period (equal to T seconds) of the signals which are of interest. To distinguish comb filters using the different types of storage system, i.e. with

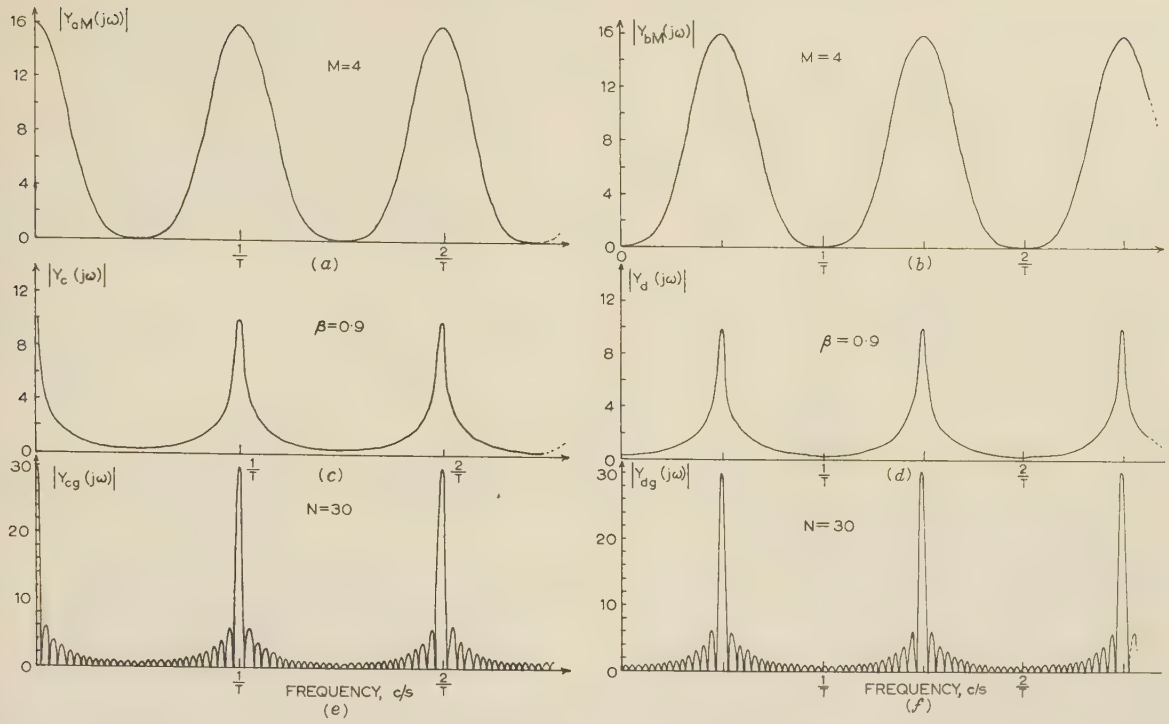


Fig. 3.—Frequency response characteristics.

continuous input and output and saturable, they will be referred to as *continuous-type* and *group-type* comb filters respectively.

The discussion of Section 4 shows that the group-type of comb filter can have the configuration of Figs. 2(*c*) and (*d*). The transfer functions for the group-type filters of these configurations can be deduced from the transfer functions for the similarly arranged continuous systems.

Suppose the group consists of N fundamental periods of duration T seconds. The transfer function for a continuous-type C is, as above,

$$Y_c(p) = \frac{\varepsilon^{-pT}}{1 - \beta\varepsilon^{-pT}} \quad . \quad . \quad . \quad . \quad . \quad (8)$$

Such a filter with continuous input can be regarded as combining together an infinite number of fundamental periods, the output remaining finite provided that $\beta < 1$. Now,

$$\frac{1}{1 - \beta\varepsilon^{-pT}} = \sum_{n=0}^{n=\infty} \beta^n \varepsilon^{-npT}$$

Thus, if N fundamental periods are combined together, the system transfer function will be

$$Y_{cg}(p) = \varepsilon^{-pT} \sum_{n=0}^{n=N-1} \beta^n \varepsilon^{-npT}$$

giving, for the transfer function of a type-C group filter,

$$Y_{cg}(p) = \varepsilon^{-pT}\left(\frac{1 - \beta^N\varepsilon^{-NpT}}{1 - \beta\varepsilon^{-pT}}\right) \quad . \quad . \quad . \quad (19)$$

when a group of N fundamental periods are combined together.

The transfer function for a continuous type-D filter is

$$Y_d(p) = \frac{\varepsilon^{-pT}}{1 + \beta\varepsilon^{-pT}} \quad . \quad . \quad . \quad . \quad . \quad (9)$$

and a similar argument gives

$$Y_d(p) = \varepsilon^{-pT} \sum_{n=0}^{n=\infty} (-\beta)^n \varepsilon^{-npT}$$

and the transfer function for a type-D group filter combining N fundamental periods together will be

$$Y_{dg}(p) = \varepsilon^{-pT} \sum_{n=0}^{n=N-1} (-\beta)^n \varepsilon^{-npT}$$

Thus $$Y_{dg}(p) = \varepsilon^{-pT}\left[\frac{1 - (-\beta)^N\varepsilon^{-NpT}}{1 + \beta\varepsilon^{-pT}}\right] \quad . \quad . \quad . \quad (20)$$

Putting $p = j\omega$ in eqns. (19) and (20), the amplitude/frequency response characteristics for these two types are obtained. For the type-C group filter,

$$Y_{cg}(j\omega) = \varepsilon^{-j\omega T}\left(\frac{1 - \beta^N\varepsilon^{-j\omega NT}}{1 - \beta\varepsilon^{-j\omega T}}\right)$$

$$= \varepsilon^{-j\omega T}\left[\frac{(1 - \beta^N \cos \omega NT) + j\beta^N \sin \omega NT}{(1 - \beta \cos \omega T) + j\beta \sin \omega T}\right]$$

whence $$|Y_{cg}(j\omega)| = \sqrt{\left(\frac{1 + \beta^{2N} - 2\beta^N \cos \omega NT}{1 + \beta^2 - 2\beta \cos \omega T}\right)} \quad . \quad (21)$$

and, similarly, for the type-D group filter,

$$|Y_{dg}(j\omega)| = \sqrt{\left[\frac{1 + \beta^{2N} - 2(-\beta)^N \cos \omega NT}{1 + \beta^2 + 2\beta \cos \omega T}\right]} \quad . \quad (22)$$

As discussed in Section 4, in the special type of filter combining the signals in groups it is possible for β to equal unity and the filter to remain stable. This would normally be done in a

group-type filter, and putting $\beta = 1$ in eqns. (21) and (22) gives the simpler expressions

$$|Y_{cg}(j\omega)| = \sqrt{\left(\frac{1 - \cos \omega NT}{1 - \cos \omega T}\right)} = \left|\frac{\sin \frac{\omega NT}{2}}{\sin \frac{\omega T}{2}}\right| \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (23)$$

and

$$|Y_{dg}(j\omega)| = \sqrt{\left[\frac{1 - (-1)^N \cos \omega NT}{1 + \cos \omega T}\right]} = \left|\frac{\cos \frac{\omega NT}{2}}{\cos \frac{\omega T}{2}}\right| \text{ or } \left|\frac{\sin \frac{\omega NT}{2}}{\cos \frac{\omega T}{2}}\right|$$

according as N is odd or even . . (24)

Amplitude/frequency response characteristics calculated from eqns. (23) and (24) are shown in Figs. 3(*e*) and (*f*) for $N = 30$.

(2.2) Frequency Response of Type A, B, C and D Filters

The comb filter is useful for the selection or rejection of repetitive signals uniformly spaced in time at an interval of T seconds, say, such signals having frequency spectra[1] centred on multiples of the basic repetition frequency, $1/T$. Inspection of the amplitude/frequency response characteristics of Fig. 3 shows that configurations in which the combining unit is subtractive tend to reject such signals, while those with additive combination tend to select them, i.e. cancellers are obtained with subtractive combining units and integrators with the additive-type combination.

Considering the cancellation filter, type B is greatly superior to type D as it gives complete rejection at the required frequencies. Thus, type B is the configuration invariably adopted for a cancellation comb filter. No greater rejection at the basic repetition frequency is obtained by cascading several units of this type, but such a procedure is often adopted to give greater selectivity and higher rejection of frequencies in the neighbourhood of the spectrum which it is desired to reject.

An integrating filter should give the highest gain possible at those frequencies corresponding to the spectrum which it is desired to select. Both types A and C are, in theory, capable of providing any required gain. In a filter of type A this necessitates cascading filters of the basic configuration shown in Fig. 2(*a*), while in type C the desired value of gain may be achieved using only one basic unit, by adjustment of β. On this basis, the type-C filter is superior to the type-A filter as an integrator, as the superior rejection performance of the latter does not normally compensate the multiplication of units necessary to achieve the required gain. Filters of type C are not normally cascaded as the gain obtained by adjustment of β normally provides sufficient selectivity.

Further discussion of the comb filter is confined to the two preferred configurations, cascaded units of type B as a canceller, and single units of type C as an integrator of both the continuous and the group type.

(2.3) Transfer Function and Frequency Response with Practical Delay Lines and Circuits

The above analysis has assumed ideal characteristics for the delay line and feedback circuit. Practical delay lines and feedback circuits will introduce amplitude distortion and phase shift, and it is important to know how this modifies the transfe function and frequency response characteristic of the com filter.

Let the transfer function of the practical delay unit be

$$Y_l(p) = \varepsilon^{-pT} Y_1(p) \quad . \quad . \quad . \quad . \quad . \quad (25$$

and that of the feedback circuit be $Y_2(p)$. The two basic arrange ments of the types B and C filters are now diagrammatically represented as in Fig. 4. The overall transfer functions of these

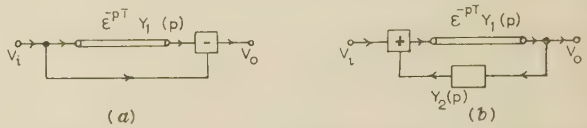


Fig. 4.—Preferred basic arrangements with practical transfer functions

circuits can be deduced by arguments exactly similar to those adopted in Section 2 for the ideal arrangements. This gives, for the type-B canceller of Fig. 4(*a*), the transfer function

$$Y_{bp}(p) = [1 - \varepsilon^{-pT} Y_1(p)] \quad . \quad . \quad . \quad . \quad (26)$$

and for the type-C integrator of Fig. 4(*b*), the transfer function

$$Y_{cp}(p) = \frac{Y_1(p)\varepsilon^{-pT}}{[1 - Y_1(p)Y_2(p)\varepsilon^{-pT}]} \quad . \quad . \quad . \quad (27)$$

These transfer functions enable the response of the overall filter to any input to be computed when the transfer functions of the individual elements are known. The effect of the characteristics of the individual units on the overall frequency response characteristic of the comb filter is discussed in Section 8.1 for the more complex case of the integrating filter. It is shown there that, provided certain reasonable assumptions can be made, the maxima and minima of the type-C integrating comb filter frequency response are given by

$$\left|Y_{cp}(j\omega)\right|_{\substack{max \\ min}} = \frac{|Y_1(j\omega)|}{[1 \mp |Y_1(j\omega)|\,|Y_2(j\omega)|]} \quad . \quad . \quad (28)$$

occurring at frequencies given by

$$\omega T = (\text{phase shift in feedback circuit}) + (\text{excess phase shift in delay unit}) + k\pi \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

The effect on the overall frequency response of the integrating comb filter is shown in Fig. 5.

Eqn. (29) shows that any phase shift in the feedback circuit or excess phase shift in the delay unit above that associated with the delay time, T, will reduce the effectiveness of the integrating filter by shifting the peaks of the response away from the optimum values given by eqn. (18). Such phase shifts must therefore be kept to the minimum, and the restrictions which this places on the characteristics of the individual units are considered in Section 4. Examination of Fig. 5, which is derived from eqn. (28), shows that the effective frequency range of operation of the comb filter is determined closely by the bandwidth of the narrowest element, delay unit or feedback circuit. Exactly similar considerations apply to the cancellation filter, whose effective range of operation is confined to the bandwidth of the delay unit and whose excess phase shift must be kept within the limits discussed in Section 4.

(3) PERFORMANCE OF COMB FILTERS

(3.1) Response to an Input-Signal Train

An input-signal train is considered consisting of a series of N regularly spaced uniform pulses at interval T seconds. Let the

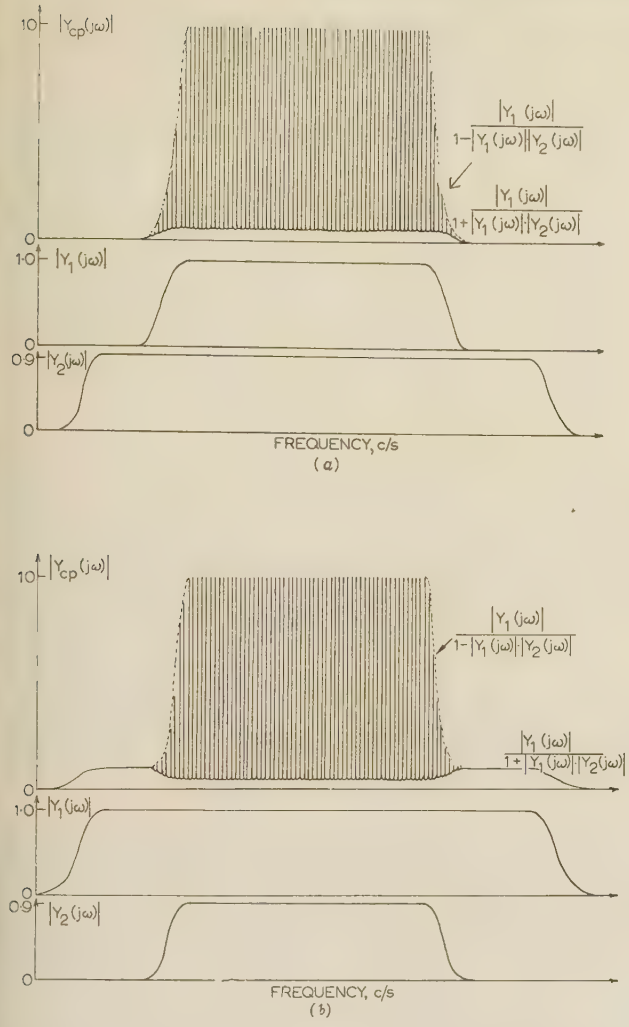


Fig. 5.—Frequency response of practical integrating filter.

first pulse, starting at time $t = 0$, be the function of time $F(t)$ having a transform $\bar{F}$. Then the transform of the second pulse, delayed by a time T, will be $\varepsilon^{-pT}\bar{F}$ and the transform of the input-signal sequence is

$$\bar{S}_i = \bar{F}[1 + \varepsilon^{-pT} + \varepsilon^{-p2T} + \ldots + \varepsilon^{-p(N-1)T}] \quad . \quad (30)$$

i.e.

$$\bar{S}_i = \bar{F}\frac{1 - \varepsilon^{-NpT}}{1 - \varepsilon^{-pT}} \quad . \quad . \quad . \quad . \quad . \quad (31)$$

(3.1.1) Signal Performance of Canceller.

The transfer function of M cascaded cancellers of the type B shown in Fig. 4(*a*) is

$$[Y_{bp}(p)]^M = [1 - \varepsilon^{-pT}Y_1(p)]^M \quad . \quad . \quad . \quad (32)$$

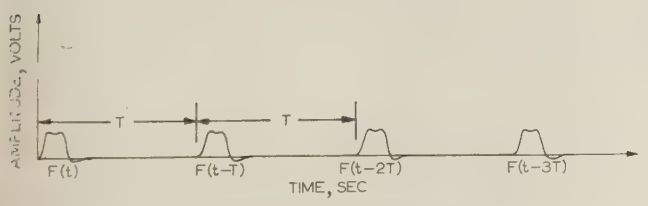


Fig. 6.—Input signal train.

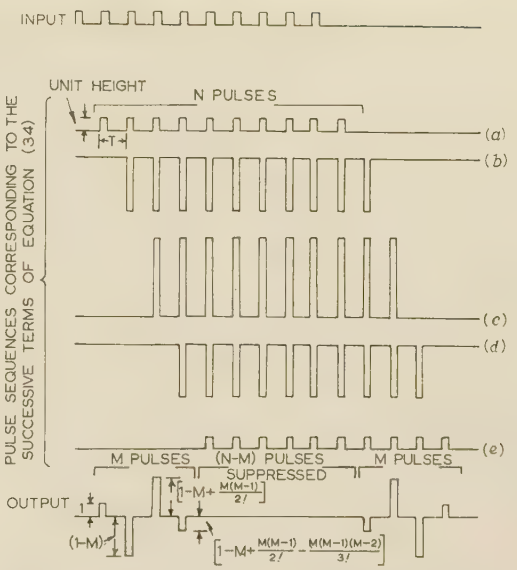


Fig. 7.—Action of ideal cascaded cancellation filter.
Drawn for $N = 10$, $M = 4$.
Output = sum of sequences (*a*) to (*e*).

and thus the output of such an M-cascaded canceller with the signal sequence of Fig. 6 fed in will be

$$S_o(t) = \mathscr{L}^{-1}[1 - \varepsilon^{-pT}Y_1(p)]^M\left(\frac{1 - \varepsilon^{-NpT}}{1 - \varepsilon^{-pT}}\right)\bar{F} \quad . \quad (33)$$

The action of this type of comb filter in suppressing a train of input pulses is made clear by considering the case of a series of M cascaded ideal cancellers fed with a train of N rectangular pulses spaced T seconds apart, as shown in Fig. 7. The expanded form of the transfer function is

$$\begin{aligned}[Y_b(p)]^M &= [1 - \varepsilon^{-pT}]^M \\ &= 1 - M\varepsilon^{-pT} + \frac{M(M-1)}{2!}\varepsilon^{-p2T} + \ldots \\ &\quad \ldots + \frac{(-1)^r M(M-1)\ldots(M-r+1)}{r!}\varepsilon^{-prT} \\ &\quad + \ldots + (-1)^M\varepsilon^{-MpT}\end{aligned}$$

Thus the transform of the filter output will be

$$\begin{aligned}\bar{S}_o = [Y_b(p)]^M\bar{S}_i &= \bar{S}_i - M\bar{S}_i\varepsilon^{-pT} + \ldots \\ &\quad \ldots + \frac{(-1)^r M(M-1)\ldots(M-r+1)}{r!}\bar{S}_i\varepsilon^{-prT} + \ldots \\ &\quad \ldots + (-1)^M\bar{S}_i\varepsilon^{-pMT}\end{aligned}$$

giving an output-signal sequence

$$\begin{aligned}S_o(t) &= S_i(t) - MS_i(t-T) + \ldots \\ &\quad \ldots + \frac{(-1)^r M(M-1)\ldots(M-r+1)}{r!}S_i(t-rT) + \ldots \\ &\quad \ldots + (-1)^M S_i(t-MT) \quad . \quad . \quad (34)\end{aligned}$$

Inspection of eqn. (34) shows that the output from such a comb filter will consist of the sum of the input pulse train and a sequence of pulse trains delayed by increasing multiples of the delay time, T, and multiplied by the appropriate binomial coefficient. This is illustrated in Fig. 7, where it can be seen that $N - M$ pulses are completely suppressed in the output train.

The magnitudes of the residual transients at the beginning and end of the output pulse train are determined by the successive

partial sums of the binomial coefficients shown in the Figure. A simple single-stage canceller fed with such an input would suppress all the pulses in the input pulse train with the exception of a single pulse at the beginning and end of the sequence. Thus, the increased selectivity of the frequency response of a cascaded canceller is achieved at the expense of an increased transient output at the beginning and end of the pulse train which it is desired to cancel. If distortion is present in the delay unit, perfect cancellation of the N–M suppressed pulses in the sequence will not be achieved, and the magnitude of the residuals may be estimated by using eqn. (33). The solution for the output transform will then contain powers of $Y_1(p)$ up to $Y_1^M(p)$, and thus distortion in the delay unit must be kept to a low level if it is proposed to cascade cancellers.

(3.1.2) **Signal Performance of Integrator.**

The transfer function of a type-C integrating comb filter as shown in Fig. 4(*b*) is

$$Y_{cp}(p) = \frac{Y_1(p)\varepsilon^{-pT}}{[1 - Y_1(p)Y_2(p)\varepsilon^{-pT}]}$$

as before. This can be expanded as

$$Y_{cp}(p) = Y_1(p)\varepsilon^{-pT}[1 + Y_1(p)Y_2(p)\varepsilon^{-pT} - Y_1^2(p)Y_2^2(p)\varepsilon^{-p2T} + \ldots]$$

Thus, for an input train of pulses as shown in Fig. 6, the transform of the output pulse sequence will be

$$\begin{aligned}\bar{S}_o &= \bar{S}_i[Y_{cp}(p)] \\ &= \bar{F}[1 + \varepsilon^{-pT} + \varepsilon^{-p2T} + \ldots + \varepsilon^{-p(N-1)T}] \\ &\quad \times [Y_1(p)\varepsilon^{-pT}][1 + Y_1(p)Y_2(p)\varepsilon^{-pT} + Y_1^2(p)Y_2^2(p)\varepsilon^{-p2T} + \ldots]\end{aligned}$$

i.e.
$$\bar{S}_o = \bar{F}Y_1(p)\{\varepsilon^{-pT} + [1 + Y_1(p)Y_2(p)]\varepsilon^{-p2T} + [1 + Y_1(p)Y_2(p) + Y_1^2(p)Y_2^2(p)]\varepsilon^{-p3T} + \ldots\} \quad (35)$$

A closed form for the coefficients of eqn. (35) is obtained in Section 8.2, but the physical action of the integrating filter is made clearer by considering the expanded form. This gives, for the filter output, the sequence

$$S_o(t) = \mathscr{L}^{-1}\bar{F}Y_1(p)\varepsilon^{-pT} + \mathscr{L}^{-1}\bar{F}Y_1(p)[1 + Y_1(p)Y_2(p)]\varepsilon^{-p2T} + \mathscr{L}^{-1}\bar{F}Y_1(p)[1 + Y_1(p)Y_2(p) + Y_1^2(p)Y_2^2(p)]\varepsilon^{-p3T} + \ldots \quad (36)$$

Eqn. (36) shows that the output consists of a sequence of pulses. The first of these pulses has been distorted and delayed by the delay unit only, the second pulse is delayed by a time $2T$ and is the sum of two pulses, one which has passed through the delay unit only, and one which has passed through the delay unit twice and the feedback circuit once, and so on. Subsequent pulses forming the output occur at all multiples of T and are formed by the summation of suitably delayed and timed pulses circulating in the feedback loop.

The action is shown in Fig. 8 for the case of an ideal integrator of type C as shown in Fig. 2(*c*) fed with a train of uniform rectangular pulses, regularly spaced at an interval T. The integrating action of the comb filter is clear and the output rises to a maximum at a time $N(T + 1)$ seconds after the first pulse is fed in. If a long sequence of pulses is fed to such an integrator, the possibility of a considerable amplitude gain for detection purposes is apparent. The magnitude of such gains is examined in Section 3.3.

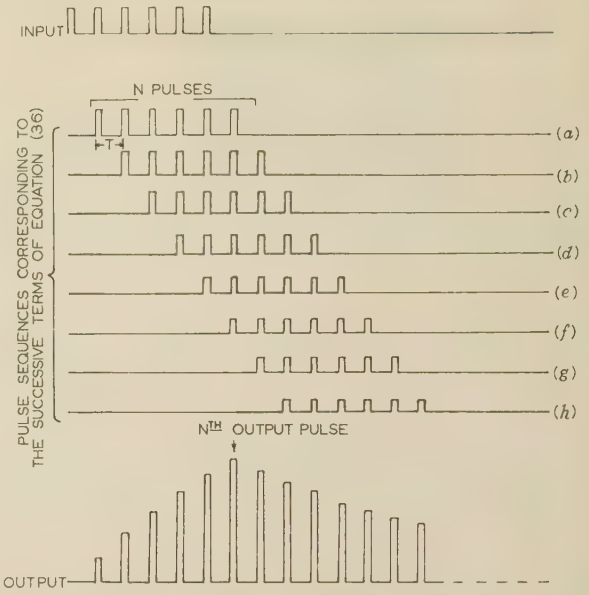


Fig. 8.—Action of ideal integrating filter.
Output = sum of sequences (*a*) + (*b*) + (*c*) + . . .

(3.2) **Response of Comb Filter to Gaussian Noise**

The statistics of a Gaussian noise waveform[13] are completely determined by its power spectrum w(f), the standard deviation being given by

$$\sigma^2 = \int_0^\infty w(f)df \quad . \quad . \quad . \quad . \quad . \quad . \quad (37)$$

Thus, the standard deviation of the output noise from an ideal comb filter fed with Gaussian noise of a known finite bandwidth can be calculated using eqn. (37) and the frequency characteristics derived in Section 2. Such a method is very laborious and it is simpler to use the expanded form of the comb filter transfer function to express the output noise waveform as the sum of a series of time-displaced noise waveforms of calculable r.m.s. value. These noise outputs can then be combined in r.m.s. fashion, ignoring the time displacements, as the statistics of a Gaussian noise waveform are stationary and the correlations are negligible. Only the noise performance of ideal filters is considered here.

(3.2.1) **Response of Canceller to Gaussian Noise.**

The expanded form of the transfer function of M cascaded cancellers of the ideal type B shown in Fig. 2(*b*) is

$$Y_b(p) = \left[1 - M\varepsilon^{-pT} + \frac{M(M-1)}{2!}\varepsilon^{-p2T} + \ldots + (-1)^M\varepsilon^{-MpT}\right]$$

and thus the output of such a filter fed with Gaussian noise will consist of a series of noise waveforms displaced in time by multiples of T seconds and having magnitudes multiplied by the appropriate binomial coefficient. For a Gaussian input, the output noise waveforms will be completely uncorrelated and the r.m.s. value of the output noise will be the square root of the sum of the squares of the individual r.m.s. values. Thus, the output r.m.s. noise is increased by a factor

$$\frac{\text{R.M.S. value of output noise}}{\text{R.M.S. value of input noise}} = \sqrt{\left[1 + M^2 + \frac{M^2(M-1)^2}{(2!)^2} + \ldots + 1^2\right]} = \frac{\sqrt{(2M)!}}{M!} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (38)$$

Eqn. (38) shows that the cascading of cancellers is penalized by a high rate of increase of the r.m.s. value of the output noise; four ideal cancellers in cascade would increase the r.m.s. value of the output noise by a factor $\sqrt{70}$, i.e. 18·46 dB, whereas the output noise of a single canceller would increase by a factor $\sqrt{2}$, i.e. 3 dB.

(3.2.2) **Response of Integrator to Gaussian Noise.**

For an ideal type-C integrating filter as shown in Fig. 2(*c*), the transfer function in its expanded form may be obtained from eqn. (27) as expanded in Section 3.1.2 by putting $Y_1(p) = 1$ and $Y_2(p) = \beta$, giving

$$Y_c(p) = \varepsilon^{-pT}(1 + \beta\varepsilon^{-pT} + \beta^2\varepsilon^{-p2T} + \ldots)$$

i.e.
$$Y_c(p) = \varepsilon^{-pT}\sum_{n=0}^{n=\infty}\beta^n\varepsilon^{-npT} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (39)$$

Inspection of eqn. (39) shows that, if such a filter is fed with a Gaussian input noise waveform, the output noise waveform can be considered as the sum of a series of noise waveforms, delayed in time by multiples of T seconds, whose amplitudes are multiplied by unity, β, β^2 and so on. Thus, if a continuous integrator is fed with Gaussian noise, the output r.m.s. value of noise is increased by a factor

$$\frac{\text{R.M.S. value of output noise}}{\text{R.M.S. value of input noise}} = \mathscr{L}t_{n\to\infty}\sqrt{(1 + \beta^2 + \beta^4 + \ldots + \beta^{2n})}$$

$$= \frac{1}{\sqrt{(1 - \beta^2)}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (40)$$

For a group integrator adding N signal sequences together, of length T, the output r.m.s. noise will be increased by a factor

$$\frac{\text{R.M.S. value of output noise}}{\text{R.M.S. value of input noise}} = \sqrt{(1 + \beta^2 + \beta^4 + \ldots + \beta^{2N-2}}$$

$$= \sqrt{\left(\frac{1 - \beta^{2N}}{1 - \beta^2}\right)} \quad . \quad . \quad . \quad . \quad (41)$$

owing to the finite number of noise waveforms combined together.

(3.3) Signal/Noise Ratio Improvement of Ideal Integrators

The performance of an ideal integrator in improving the signal/noise ratio of a train of uniformly spaced pulses of rectangular form embedded in Gaussian noise is now investigated. The detection problem to which this discussion is relevant is the simple one of deciding whether the pulse train is present or absent. The probability of successful detection (i.e. a correct guess) is a function of the signal/noise ratio. For the purposes of this discussion we define the signal/noise ratio as the numerical ratio of the peak amplitude of the largest pulse in the train being examined to the r.m.s. value of the noise.

(3.3.1) **Continuous Integrator.**

The action of the filter in increasing the peak amplitude of the pulses in the output train is made clearer by Fig. 8. Suppose the peak value of the pulses in the input train is V and that the r.m.s. value of the input noise is V_1, an input signal/noise ratio of V/V_1. For a train of N input pulses,

$$\text{Peak value of } N\text{th output pulse} = \sum_{n=0}^{n=N-1}\beta^n V = V\left(\frac{1 - \beta^N}{1 - \beta}\right)$$

$$\text{R.M.S. value of output noise} = \frac{V_1}{\sqrt{(1 - \beta^2)}}$$

Therefore
$$\frac{\text{Output signal}}{\text{Output noise}} = \frac{V\left(\frac{1 - \beta^N}{1 - \beta}\right)}{\frac{V_1}{\sqrt{(1 - \beta^2)}}}$$

If the ratio of output signal/noise to input signal/noise is termed the signal/noise ratio improvement of the integrator, I_{SN}, then

$$I_{SN} = \left(\frac{1 - \beta^N}{1 - \beta}\right)\sqrt{(1 - \beta^2)}$$

$$= (1 - \beta^N)\sqrt{\left(\frac{1 + \beta}{1 - \beta}\right)} \quad . \quad . \quad . \quad . \quad (43)$$

The graph of eqn. (43) for a series of values of N is shown in Fig. 9, where it is seen that, for $N = 100$, the signal/noise ratio

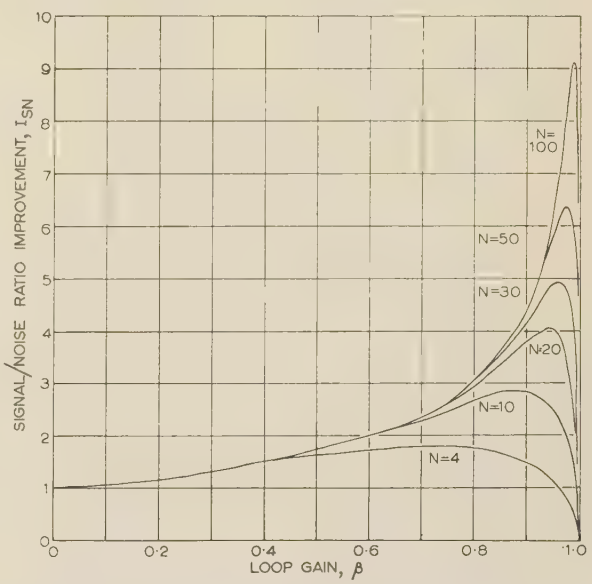


Fig. 9.—Signal/noise ratio improvement of ideal continuous integrator.

is improved by a factor of 9·1, or 19·18 dB, at the optimum value of β. The data presented in Fig. 9 show that there is an optimum value of β for any given number of input pulses, which must be closely maintained in order to achieve the inherent signal/noise ratio improvement of the integrator. The implications of this in considering the design of such a filter are discussed in Section 4.

(3.3.2) **Group Integrator.**

The signal/noise ratio improvement of this type of integrator differs from that of the continuous integrator, because the noise is not continuously integrated but is combined as a finite group of noise waveforms of duration T, in number equal to the number of signal pulses being combined together. The peak value of the Nth output pulse will be the same as in the continuous case, i.e. $V[(1 - \beta^N)/(1 - \beta)]$. If the input noise has an r.m.s. value V_1, the output noise will have an r.m.s. value given by

R.M.S. value of output noise

$$= V_1\sqrt{\left[\frac{(1 - \beta^{2N})}{(1 - \beta^2)}\right]} = V_1\sqrt{\left[\frac{(1 - \beta^N)(1 + \beta^N)}{(1 - \beta)(1 - \beta)}\right]}$$

as in Section (3.2.2). Thus, the signal/noise ratio improvement of an ideal group type integrating filter will be

$$I_{SN} = \frac{\dfrac{1-\beta^N}{1-\beta}}{\sqrt{\left[\dfrac{(1-\beta^N)(1+\beta^N)}{(1-\beta)(1+\beta)}\right]}} = \sqrt{\left[\frac{(1+\beta)(1-\beta^N)}{(1-\beta)(1+\beta^N)}\right]} \quad . \quad (44)$$

and the graph of this equation for a series of values of N is given in Fig. 10. It is seen that, for the group-type of integrator, the

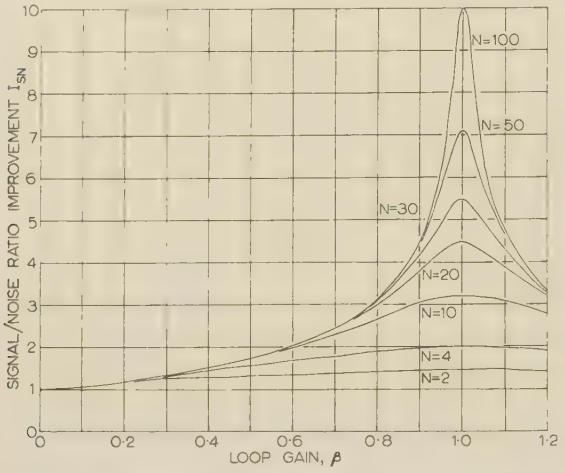


Fig. 10.—Signal/noise ratio improvement of group integrator.

optimum value of β is always unity. It is shown in Section 8.3 that the signal/noise ratio improvement when β is unity becomes

$$I_{SN\,max} = \sqrt{N} \quad . \quad . \quad . \quad . \quad . \quad . \quad (45)$$

i.e. a factor of 3 dB per doubling of the number of pulses added together.

(4) PRACTICAL FORMS OF COMB FILTER

The discussion in the paper has been confined to the type of comb filter which uses the combinations of delay and combining unit shown in Fig. 1. The practical form taken by such a device is largely determined by the type of delay or storage unit used. The alternative approach of paralleling a large number of single-peaked response filters seems applicable only when a very simple form of comb filter response is required with, say, up to 100 response peaks. Such filters have been constructed[20] and normally use magnetostrictive elements, quartz crystals, vibrating strings or other electro-mechanical devices to obtain the narrow bandwidths required, which are of the order of 1–10 c/s. (The width between 3 dB points of the teeth of the response of the type of comb filter considered here may be estimated as in Section 8.4.)

If a practical system of the type discussed here is to operate efficiently, certain stability requirements (in the sense of lack of variation of defining parameters with time) must be satisfied. These stability requirements may be deduced from the frequency response characteristics of the system. Consideration of the main results of the above analysis, together with these stability requirements, then indicates the form which the practical realization of cancellation and integration filters must take.

(4.1) Stability Requirements

The input signal is assumed to be a train of uniformly spaced pulses at intervals, T. In theory, a comb filter may be built to operate at any frequency for which a storage unit is available. Suppose that the input pulse train consists of a modulated carrier of frequency f_c and that the modulating pulses have an effective bandwidth, B. The input signal spectrum will be centred on f_c with peaks spaced on either side at multiples of $1/T$ and contained within limits $f_c \pm B$. Fig. 11 shows the

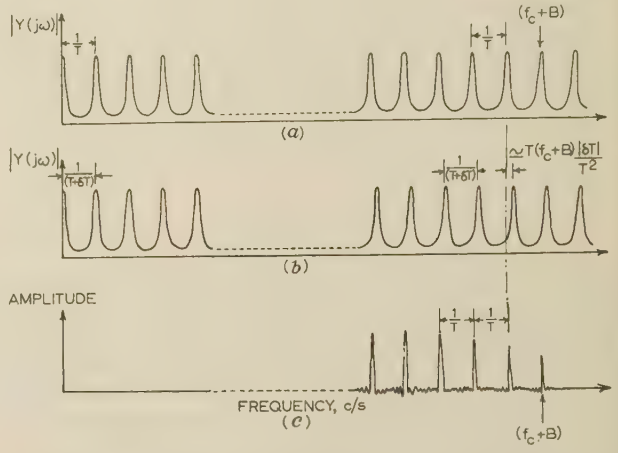


Fig. 11.—Effect of change in delay-unit delay time.
(*a*) Frequency response of integrating comb filter with delay time T sec.
(*b*) The same, with delay time $(T + \delta T)$ sec.
(*c*) Frequency spectrum of input pulse train.

effect of a change in delay-unit delay time on the response of an integrating type of filter. The delay-unit delay time is assumed to change from T to $T + \delta T$. Let $1/T = f$, then

$$\frac{1}{T+\delta T} = f + \delta f$$

where δf will be the frequency shift of the first peak of the comb filter response.

Thus
$$f + \delta f \simeq \frac{1}{T}\left(1 - \frac{\delta T}{T}\right)$$

i.e.
$$\delta f \simeq \frac{|\delta T|}{T^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (46)$$

The number of response peaks included in the comb filter response up to the highest frequency of interest, $f_c + B$, is approximately $T(f_c + B)$. Suppose that the permissible shift in the frequency response peak enclosing the signal-spectrum peak at $(f_c + B)$ is b. Then, to a first order of approximation, the permissible change in line delay time will be given by

$$T(f_c + B)\frac{|\delta T|}{T^2} < b$$

i.e.
$$|\delta T| < \frac{bT}{f_c + B} \quad . \quad . \quad . \quad . \quad . \quad (47)$$

The maximum permissible shift parameter b is discussed in Section 8.4. The orders of magnitude involved may be illustrated by considering the case of a continuous integrator with an input train of 20 pulses. Fig. 9 shows that the optimum value of β is 0·95. Assuming that the frequency shift of the highest response peak of interest is to be restricted so that the spectral line at $f_c + B$ falls to the 3 dB point of the relevant response peak and that T is 4 millisec, the method of Section 8·4 gives the permissible value of b as 2 c/s. Thus, if $f_c = 30$ Mc/s and $B = 100$ kc/s, then

$$|\delta T| < \frac{2 \times 4 \times 10^{-3}}{30{\cdot}1 \times 10^{5}} \text{ sec}$$

i.e. $$|\delta T| < 0{\cdot}266 \times 10^{-9}\,\text{sec}$$

The presence of f_c in eqn. (47) places such stringent restrictions on the stability of the delay time of the delay element that the construction of comb filters to operate at other than video frequencies and lower does not seem feasible. A filter similar to the above, but operating only on the modulating pulses themselves, no carrier being involved, would require

$$|\delta T| < \frac{2 \times 4 \times 10^{-3}}{10^5}\,\text{sec}$$

i.e. $$|\delta T| < 0{\cdot}08\,\text{microsec}$$

an order of stability attainable in practice.

(4.2) Bandwidth Requirements

The bandwidth of the individual elements of the comb filter must be sufficient to keep the distortion of individual pulses being handled by the filter within prescribed limits in order that it may function as intended. The frequency transfer function of an M-cascaded canceller is obtained from eqn. (32) as

$$[Y_{bp}(j\omega)]^M = [1 - \varepsilon^{-j\omega T}Y_1(j\omega)]^M \quad . \quad . \quad . \quad (48)$$

and its expanded form will contain terms up to the degree of $Y_1^M(j\omega)$. Thus the bandwidth of the delay element response must be such that the bandwidth of $Y_1^M(j\omega)$ is comparable with that of the frequency spectrum of an individual input pulse. As M is rarely greater than 2 or 3 owing to the resulting increased noise and transient output, no bandwidth difficulties will be encountered in this type of filter if a quartz delay line is used, until the input-pulse spectrum bandwidth approaches 5 Mc/s.

The restrictions on the bandwidth of the elements used in an integrating comb filter are, however, much more severe. Its transfer function is such that eqn. (36) gives the spectrum of the Nth pulse in the output sequence as

$$F_N(j\omega) = F(j\omega)Y_1(j\omega)$$
$$[1 + Y_1(j\omega)Y_2(j\omega) + \ldots + Y_1^{N-1}(j\omega)Y_2^{N-1}(j\omega)]$$

With a large number of input pulses, say 100, the highest-order distortion term must be such that its bandwidth is comparable with that of an input pulse. As the term in question is $Y_1^N(j\omega)Y_2^{N-1}(j\omega)$, this places severe restrictions on the individual bandwidths of $Y_1(j\omega)$ and $Y_2(j\omega)$. The orders of magnitude involved may be estimated by assuming the 3 dB reduction factor to be $\sqrt{(2^{1/N}-1)}$ as in cascaded video- or single-tuned radio-frequency stages. If $N = 100$, this reduction factor is 0·083 and thus an input bandwidth of 1 Mc/s in the input-pulse spectrum would demand a bandwidth of the order of 12 Mc/s in $Y_1(j\omega)Y_2(j\omega)$, which would be very difficult to achieve in practice in the storage system.

(4.3) Phase Shift Requirements

Section 8.1 shows that in a practical integrating comb filter, the response peaks occur at frequencies given by

$$\omega T = (\text{phase shift in feedback circuit})$$
$$+ (\text{excess phase shift in delay unit}) + k\pi$$

and thus the phase shift in any circuits associated with the delay line and with the feedback circuit should be kept as small as possible. The amount of phase shift in any circuits associated with the units is related to the flatness of their frequency response characteristics in the operating frequency region. The relationship between amplitude response and phase shift is unique for a minimum-phase-shift type network, and for a feedback network of this type an estimate can be made of the restrictions on amplitude variations of the frequency response characteristics over the pass-band. The delay unit is not of minimum-phase-shift type and no unique criterion obtains; excess phase shift may be introduced through ancillary apparatus feeding the delay unit and it would seem reasonable to assume that the same orders of restriction on the flatness of the amplitude response deduced for the feedback unit can be applied to the delay unit as a whole.

If, over an interval of several octaves, the amplitude response has a constant slope of K dB per octave, the associated phase shift is given by[19]

$$\theta = \frac{k\pi}{12}\,\text{radians} \quad . \quad . \quad . \quad . \quad . \quad (49)$$

Let the permissible frequency shift of the highest response peak of interest be b, and assume that θ is the only phase shift occurring in the system. Then, from eqn. (29), $\omega T = k\pi + \theta$ and response peaks occur at frequencies given by

$$\omega = \frac{k\pi}{T} + \frac{\theta}{T}$$

Thus $$\frac{\theta}{T} < 2\pi b$$

i.e. $$\frac{k\pi}{12}\,\frac{1}{T} < 2\pi b$$

or $$K < 24bT \quad . \quad . \quad . \quad . \quad . \quad . \quad (50)$$

The orders of magnitude involved may be seen by taking $T = 4$ millisec and $b = 2$ c/s, which gives $K < 0{\cdot}192$ dB per octave.

(4.4) Available Storage Media

Available storage elements[9] may be broadly classified under two headings, static and dynamic. The dynamic elements utilize propagation in a medium at a velocity considerably less than that of light, and the static elements involve the impression on some medium of a static pattern persisting in time. Dynamic storage is available in distributed and lumped delay lines and in ultrasonic delay lines in quartz, mercury and wire. Static storage (in the sense defined above) is available in the magnetic drum, magnetic tape and wire, the magnetic matrix and the electrostatic storage tube. In certain of these types of storage, the limited dynamic range available is such that the information must be encoded,[2] usually in binary digit form, before storage; this applies to ultrasonic propagation in wire and to the magnetic matrix. All the above types of storage operate with video-frequency input and output, with the exception of ultrasonic quartz and mercury delay lines, which operate with ultrasonic carriers of the order of 20–30 Mc/s. When used in video-frequency filters, such ultrasonic carrier systems are modulated with the video frequency at input and demodulated at output, all the combining and feedback circuits of the comb filter operating at video frequency. Maximum information bandwidths available are of the order of 5–10 Mc/s for the ultrasonic quartz and mercury lines, 100 kc/s for ultrasonic wire stores of between 2 and 4 millisec delay time and of the order of 1 Mc/s for all other types of storage.

(4.5) Practical Cancellers

The best configuration for a cancellation filter among the arrangements discussed here is that of type B, Fig. 2(*b*). Perfect cancellation can be obtained only if the frequency spectrum which

it is desired to suppress consists of discrete spectral lines spaced at frequency intervals of $1/T$. Such a spectrum corresponds to an infinite sequence of unvarying pulses, whereas in practice the input pulse group will have a spread of frequency components centred on multiples of $1/T$. The rejection of frequencies in the neighbourhood of multiples of $1/T$ can be increased by cascading filters of type B, as shown in Section 2. The analysis of Section 3, however, shows that such cascading results in increased transient output at the beginning and end of the pulse train which it is desired to suppress, together with a rapid increase in the output r.m.s. noise level if the input noise has a Gaussian amplitude distribution. It is thus unlikely that any advantage will be gained by cascading more than two or three basic cancellers of type B, individual calculation being necessary to resolve the relative merits and disadvantages of a particular case. Bandwidth reduction factors to be considered are small and, unlike the integrator in Section 4.6, there is no advantage to be gained by adopting a complex storage system.

The determining features in a practical design are the delay time and bandwidth required in the storage unit. An ultrasonic delay line of mercury[11] or quartz offers the best solution to the problem for delays ranging from a few microseconds up to 2 millisec with bandwidths of up to 10 Mc/s. Longer delays may be obtained by cascading delay lines, but difficulty may be experienced if there is a high attenuation in the type of ultrasonic line being used. Delays of more than a few milliseconds are best obtained by using electro-mechanical means, such as spaced recording and output heads on a magnetic drum or tape system.

A block diagram of a practical cancellation system[23] using an ultrasonic line is shown in Fig. 12. Several points must be considered if efficient pulse cancellation is to be achieved. Both delayed and undelayed pulses should be subject to attenuation and amplification in channels whose transmission parameters are as closely matched as possible. In particular, the net gains of both channels should be identical and normally the gain of one channel is automatically controlled to ensure this. The delay time of an ultrasonic delay line is subject to change with variations in ambient temperature, and steps must be taken, either to vary the system pulse-repetition frequency accordingly, or to hold the net delay time constant. If the system pulse-repetition frequency can be independently controlled, the best solution is to use the delay line directly to control the system pulse-repetition frequency by means of an interlaced reference pulse, as indicated in Fig. 12. A system with an independently fixed pulse-repetition rate must use either a temperature-controlled container for the delay line or a controlled variable compensating delay line.[11]

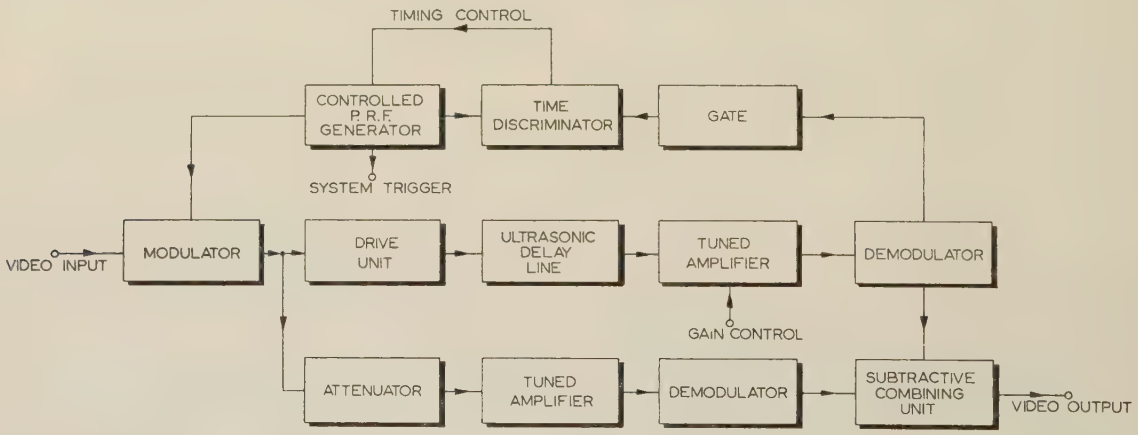


Fig. 12.—Cancellation system using ultrasonic delay line.

(4.6) Practical Integrators

The best configuration for an integrating filter among those discussed is that of type C, Fig. 2(*c*). Such an integrating filter may be either of the continuous-input or group-handling type. The continuous-input integrating filter has several disadvantages, apparent from an examination of Fig. 9. The optimum value of the loop gain, β, varies with the number of pulses being combined, and the signal/noise ratio improvement of the filter is critically dependent on this value of β. With a large number of input pulses the optimum value of the loop gain approaches unity very closely, e.g. with $N = 100$ the optimum value of β is 0·98, giving a loop gain in the positive-feedback loop of only 2% below unity. In systems handling a large number of pulses, the high and critical value of β required raises severe problems of stabilization and it is then a matter of extreme difficulty to realize the full inherent improvement of a type-C integrating filter. Systems of this type have been constructed to handle moderate numbers of pulses, and improvements of up to 4 (12 dB) have been quoted.[11] A block diagram of a continuous-input integrator using an ultrasonic delay line is shown in Fig. 13.

The adoption of a group-handling technique offers several important advantages. Examination of Fig. 10 shows that this type of filter has an optimum loop gain of unity for any number of input pulses and gives a higher value of signal/noise ratio improvement than does the continuous-integrating filter with optimum β. The group integrator essentially combines together a group of basic repetition periods of signal-plus-noise input,

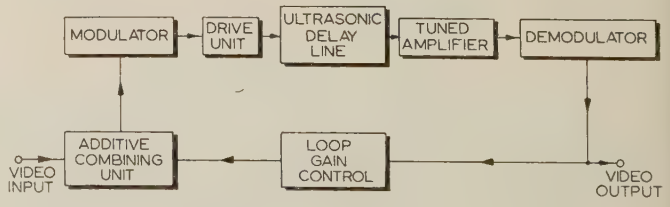


Fig. 13.—Integration system using ultrasonic delay line.

and it is possible, by using straightforward summation techniques and dispensing with the feedback loop, to produce unity-gain combination schemes which are perfectly stable. Fig. 14 shows a simple possible scheme using a number of delay lines equal to the number of basic repetition periods being combined together,

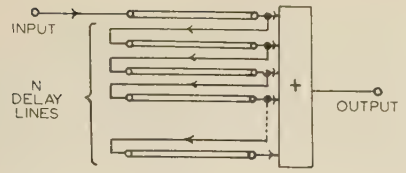


Fig. 14.—Group-handling combination scheme.

all feeding into a multiple-input combining unit. The exact adoption of such a scheme would be extravagant in the use of delay lines, but the same result can be achieved in a number of ways with greater economy if the combination is made an integral part of the storage system, giving storage of the cumulative sum of the input groups.

A well-known approach to the problem of cumulative group storage is the use of an electrostatic storage tube.[5, 7, 8, 10] Fig. 15

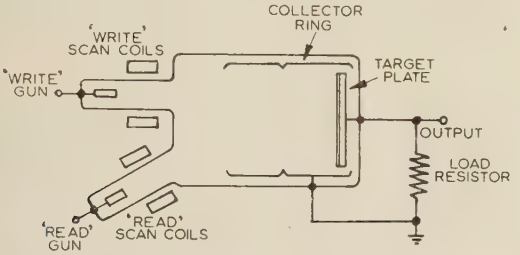


Fig. 15.—Electrostatic storage tube.

shows a type of storage tube which may be used.[5] The time variations of the input signal are recorded as a pattern of electrostatic charge on the dielectric surface of the target plate, the magnitude of the charge being linearly proportional to the input signal over a range of about 20 dB of input-signal amplitude. The 'write' beam is scanned over the surface of the target plate and operated at such a voltage (about 8 kV) that the target surface charges negatively. Repeated scanning of the 'write' beam in synchronism with the input-signal groups gives a charge distribution proportional to the cumulative sum of the input-signal traces, thus effectively integrating the input-signal groups by direct addition of charges on the target surface. After the required number of traces has been integrated, the 'read' beam is scanned over the target surface; small changes of surface charge modulate the 'read' beam current in such a way that, if the surface is negatively charged, there is an increase of secondary-emission current to the secondary-emission collector ring. These changes in secondary-emission current give an output voltage across the load resistor connected between target and collector ring. This voltage when amplified is the output of the integration system. Systems using forms of electrostatic storage tube suffer from the disadvantages of having a poor resolution and a linear range limited to about 20 dB of output-signal amplitude.

The realization of an integrating system capable of achieving the optimum signal/noise ratio improvement of $\sqrt{N}$ for N input pulses when N becomes large, say of the order of a hundred pulses, demands the use of elaborate storage and combining systems. The problems raised by the repeated operations on signals being stored and operated on in an integration system using a propagating type of storage medium are essentially similar to those involved in long-distance telephonic transmission,[21] as each involves the attenuation and amplification of the information being handled a large number of times. If an electrostatic storage system, or any other type of static storage, cannot supply the necessary resolution and dynamic range required, a propagating type of storage must be used, and this may place excessive restrictions on the bandwidth and noise properties of the system components. These restrictions may be overcome, at the cost of greatly increased complexity, by the adoption of a form of encoding and decoding[2, 15] of the signals before and after storage. The most economical systems in terms of storage capacity, as indicated above, store a cumulative sum, and, under certain conditions when large numbers of pulses are handled, an attractive solution is to encode the information prior to storage in binary digit form and perform the additive combination direct on the stored data in digital form, thus storing the cumulative digital sum which may be decoded as the output when the required number of traces has been integrated. A block diagram of such a system would be as shown in Fig. 16.

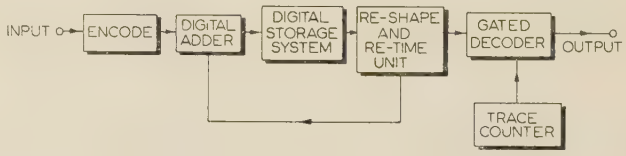


Fig. 16.—Digital integration system.

For the storage of information in simple binary form, an ultrasonic nickel line[12] may be used, which is much cheaper and more compact than a quartz or mercury line.

(4.7) The Incorporation of Comb Filters in Radar Systems

The detection of wanted targets at the output of a radar receiving system may be rendered difficult by the presence of unwanted returns (usually referred to as clutter),[23, 26, 28] and the ultimate range at which a given target can be detected will always be determined by the output level of the receiving-system noise.[23, 26] In the particular case where the obscuring clutter consists of ground or other slowly-varying returns, a canceller arrangement can be used to largely remove such stationary responses.[23] If a sufficient number of transmitter pulses are transmitted during the time that the radar aerial beam passes over the target, an integrating system may be used to improve the signal/noise ratio of the received signal returns at long ranges.[6]

(4.7.1) Incorporation of Canceller.

A cancellation scheme may be adopted when the desired responses are those from moving targets.[23, 11] The Doppler shift in the return frequency from a moving target gives rise to a beating signal when such returns are mixed with a reference frequency derived from the transmitter frequency. Such fluctuating pulse returns pass through the cancellation system and are thus separated from stationary returns, which are cancelled. Radar systems of this sort are normally classified as coherent or incoherent; coherent systems use coherent detection[16, 17, 18] with a reference-frequency voltage derived from the transmitter frequency, while incoherent systems rely on transmitter-frequency returns from stationary objects to beat with the moving-target returns and have a normal mixer and amplitude detection system. The coherent system has the great advantage of detecting moving targets, both in the presence and the absence of clutter.

A major disadvantage of any moving-target indication system working with a fixed pulse-repetition rate is the presence of target blind speeds[23, 22] at which target returns fluctuating at multiples of the pulse-repetition rate are cancelled. These blind speeds may be removed by the use of complex systems which periodically change repetition rate and matching-line delay time simultaneously.[22] The performance of a moving-target indica-

tion system using a comb filter is chiefly determined by the fluctuations present in the actual clutter return,[23] principally due to aerial scanning, internal clutter fluctuations and instability of the transmitter and cancellation systems. These fluctuations set a limit to the possible degree of cancellation of clutter. Signals from moving targets have been detected at power levels 46 dB below obscuring clutter.[27]

(4.7.2) **Incorporation of Integrator.**

An integration system may be adopted when a sufficient number of pulses is returned from the target to make the possible signal/noise ratio improvement worth the complexity introduced.[6] A major consideration in the overall design of a radar system in which it is desired to incorporate an integrating filter is the type of receiving-system demodulator used. Coherent demodulation gives a superior performance in the demodulation of signals below noise level[2, 17, 18] and is to be preferred if the additional complexity of a fully coherent system can be incorporated.

The use of an integrating filter in a system designed for the detection of signals in an obscuring noise background involves the decision whether the signal sought is present or absent.[24, 26] This decision may be left to an operator viewing a final display, or a comparison may be made automatically with some reference level before display. The use of a normal cathode-ray-tube display for inspection of the integrated signal is not desirable owing to its limited dynamic range, and it is normal to effect a comparison with a pre-set threshold level using a simple diode level-selector before display. The performance of such a system, using a square-law detector and an averaging process equivalent to an ideal group integrator of the type discussed here, has been investigated by Kaplan[24] and results have been derived showing the relationship between the signal/noise ratio of the returns before integration, the number of pulses, the proportion of valid detections with signal present and the proportion of false detections when signal is absent.

(5) CONCLUSIONS

The basic arrangements of delay unit, combining unit and feedback unit shown in Fig. 2 are seen to have amplitude/frequency response characteristics of the type which has given rise to the name comb filter. Such filters are useful for the selection or rejection of a train of uniformly-spaced regular pulses. The most suitable configuration for a cancellation filter is that of Fig. 2(*b*). Such a simple canceller will remove all pulses in such an input train, with the exception of one pulse at the beginning and one at the end of the sequence, and will double the power level of the output noise if the input noise has a Gaussian amplitude distribution. The cascading of filters of this type produces an increased transient output at the beginning and end of the pulse sequence and a rapid increase in the output noise level, given by eqn. (38). The most suitable configuration for an integrating filter of the continuous type is that of Fig. 2(*c*), having a signal/noise ratio improvement performance given by Fig. 9. Such a type of filter, with continuous input and output, is unsatisfactory for large numbers of pulses, and the best method of realizing the full inherent gain of such a scheme is to use a group-combination scheme with an effective gain of unity in the combining unit; such a scheme will give an optimum signal/noise ratio improvement of $\sqrt{N}$ for N input pulses.

Practical cancellation filters for inputs having pulse intervals of up to a few milliseconds are best realized by the use of an ultrasonic delay line unit in an arrangement similar to that of Fig. 12. For a system integrating a large number of pulses, the realization of an integrating filter of the group type demands complex storage and combining units. If the available dynamic range is sufficient, low-resolution systems can be effectively achieved using a form of electrostatic storage tube. The ultimate realization of a high-resolution large-dynamic-range wide-band system seems best achieved by the use of a simple form of propagating storage having input and output and combination in coded form, a binary digit scheme being the most suitable.

(6) ACKNOWLEDGMENTS

The author is grateful to Mr. K. Jones of the Metropolitan-Vickers Electrical Co., Ltd., who helped with much of the laborious computation involved in drawing Figs. 9 and 10 and verified some of the analytical results of Sections 2 and 3.

(7) REFERENCES

(1) George, S. F., and Zamanakos, A. S.: 'Comb Filters for Pulsed Radar Use', *Proceedings of the Institute of Radio Engineers*, 1954, **42**, p. 1159.
(2) Bell, D. A.: 'Information Theory and its Engineering Applications' (Pitman, 1953).
(3) Jaeger, J. C.: 'An Introduction to the Laplace Transformation' (Methuen, 1949).
(4) Urkowitz, H.: 'Analysis and Synthesis of Delay Line Periodic Filters', *Transactions of the Institute of Radio Engineers*, 1957, **CT–4**, p. 41.
(5) Waters, P. L.: 'Video Signal Integration Using a Storage Tube', Symposium on Communication Theory, 1953, Paper No. 17, p. 236 (Butterworths Scientific Publications).
(6) Bailey, A. E.: 'Integration in Pulse Radar Systems', *ibid.*, Paper No. 15, p. 216.
(7) Kramer, A. S.: 'Cathode Ray Storage Tubes for Special Purposes', *Electronics*, 30th January, 1959, **32**, p. 52.
(8) Harrington, J. V., and Rogers, T. F.: 'Signal-to-Noise Improvement through Integration in a Storage Tube', *Proceedings of the Institute of Radio Engineers*, 1950, **38**, p. 1197.
(9) Hollingdale, S. H.: 'High Speed Computing Methods and Applications' (English Universities Press, 1959).
(10) Thomson, J., and Callick, E. B.: 'Electron Physics and Technology' (English Universities Press, 1959).
(11) Brockelsby, C. F.: 'Ultrasonic Mercury Delay Lines', *Electronic and Radio Engineer*, 1958, **35**, p. 446.
(12) Scarrott, G. G., and Naylor, R.: 'Wire-type Acoustic Delay Lines for Digital Storage', *Proceedings I.E.E.*, Paper No. 2027 M, March, 1956 (**103 B**, Suppl. No. 3, p. 497).
(13) Rice, S. O.: 'A First Look at Random Noise', *Transactions of the American I.E.E.*, 1956, **75**, Part I, p. 128.
(14) Van Der Zeil, A.: 'Noise' (Prentice-Hall, 1955).
(15) Mayer, H. F.: 'Principles of Pulse Code Modulation', *Advances in Electronics*, 1951, **3**, p. 221.
(16) Kitai, R.: 'Coherent and Incoherent Detectors', *Electronic and Radio Engineer*, 1957, **34**, p. 96.
(17) Smith, R. A.: 'The Relative Advantages of Coherent and Incoherent Detectors: A Study of their Output Noise Spectra under Various Conditions', *Proceedings I.E.E.*, Monograph No. 6, August, 1951, **98**, Part IV, p. 43.
(18) Bell, D. A.: 'Demodulation and Detection', *Electronic and Radio Engineer*, 1959, **36**, p. 21.
(19) West, J. C.: 'Textbook of Servomechanisms' (English Universities Press, 1953).
(20) Thiele, A. P.: 'Narrow Band Magnetostrictive Filters', *Electronic and Radio Engineering*, 1958, **35**, p. 402.

(21) MEACHAM, L. A., and PETERSON, E.: 'An Experimental Multi-Channel Pulse Code Modulation System of Toll Quality', *Bell System Technical Journal*, 1948, **27**, p. 1.
(22) PERLMAN, S. E.: 'Staggered Rep Rate Fills Radar Blind Spots', *Electronics*, 21st November, 1958, **31**, p. 82.
(23) RIDENOUR, L. N.: 'Radar System Engineering', M.I.T. Radiation Laboratory Series, Vol. I (McGraw-Hill, 1947).
(24) KAPLAN, E. L.: 'Signal-Detection Studies with Applications', *Bell System Technical Journal*, 1955, **34**, p. 403.
(25) VAN VOORHIS, S. N.: 'Microwave Receivers', M.I.T. Radiation Laboratory Series, Vol. 23 (McGraw-Hill, 1948).
(26) LAWSON, J. L., and UHLENBECK, G. E.: 'Threshold Signals', M.I.T. Radiation Laboratory Series, Vol. 24 (McGraw-Hill, 1950).
(27) 'Airfield Radar Developments', *Wireless World*, 1955, **61**, p. 243.
(28) CRONEY, J.: 'Clutter on Radar Displays', *Wireless Engineer*, 1956, **33**, p. 83.

(8) APPENDICES

(8.1) Frequency Response Characteristics of Practical Comb Filters

The frequency response characteristics of the circuits of Figs. 4(*a*) and (*b*) can be deduced from eqns. (26) and (27) of Section 2.3 by putting $p = j\omega$. Considering the more complex case of the integrator, this gives

$$Y_{cp}(j\omega) = \frac{Y_1(j\omega)\varepsilon^{-j\omega T}}{[1 - Y_1(j\omega)Y_2(j\omega)\varepsilon^{-j\omega T}]} = \frac{F_1(j\omega)}{F_2(j\omega)} \text{ say.}$$

Suppose $Y_1(j\omega) = A(\omega) + jB(\omega)$

and $Y_2(j\omega) = C(\omega) + jD(\omega)$

then the denominator will be

$$\begin{aligned} F_2(j\omega) &= 1 - [A(\omega) + jB(\omega)][C(\omega) + jD(\omega)](\cos\omega T - j\sin\omega T) \\ &= 1 - \{[A(\omega)C(\omega) - B(\omega)D(\omega)] \\ &\quad + j[A(\omega)D(\omega) + B(\omega)C(\omega)]\}(\cos\omega T - j\sin\omega T) \\ &= 1 - [\alpha(\omega) + j\beta(\omega)](\cos\omega T - j\sin\omega T) \text{ say} \end{aligned}$$

giving

$$F_2(j\omega) = \{1 - [\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T]\} + j[\alpha(\omega)\sin\omega T - \beta(\omega)\cos\omega T]$$

Thus

$$\begin{aligned} |F_2(j\omega)|^2 &= 1 - 2[\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T] \\ &\quad + [\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T]^2 \\ &\quad + [\alpha(\omega)\sin\omega T - \beta(\omega)\cos\omega T]^2 \\ &= 1 + \alpha^2(\omega) + \beta^2(\omega) - 2[\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T] \end{aligned}$$

Now, $\alpha^2(\omega) + \beta^2(\omega) = [A(\omega)C(\omega) - B(\omega)D(\omega)]^2 + [A(\omega)D(\omega) + B(\omega)C(\omega)]^2$

which simplifies to

$$= [A^2(\omega) + B^2(\omega)][C^2(\omega) + D^2(\omega)]$$

i.e. $\alpha^2(\omega) + \beta^2(\omega) = |Y_1(j\omega)|^2|Y_2(j\omega)|^2$

Thus, as $|\varepsilon^{-j\omega T}| = 1$

$$|Y_{cp}(j\omega)| = \frac{|Y_1(j\omega)|}{\sqrt{\{1 + |Y_1(j\omega)|^2|Y_2(j\omega)|^2 - 2[\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T]\}}} \quad \ldots \quad (51)$$

The detailed structure of the frequency response $|Y_{cp}(j\omega)|$ can now be examined by determining the maxima and minima and the frequencies at which they occur. In practical filters of the type contemplated in the paper, $|Y_1(j\omega)|$ and $|Y_2(j\omega)|$ will have bandwidths of the order of 1 Mc/s, say, while T, the delay time, will be of the order of 1 millisec, i.e. $\cos\omega T$ and $\sin\omega T$ complete one cycle of values as the input frequency is varied over an interval of the order of 1 000 c/s. Thus, $|Y_1(j\omega)|$, $|Y_2(j\omega)|$, $\alpha(\omega)$ and $\beta(\omega)$ all vary slowly with respect to ω when compared with the rate of variation of $\cos\omega T$ and $\sin\omega T$ and can thus be regarded as constant over any small region of interest.

$|Y_{cp}(j\omega)|$ will have maxima and minima at the minima and maxima of $|F_2(j\omega)|$ if $|F_1(j\omega)|$ is regarded as constant over the range of frequencies being investigated. Thus, treating $|Y_1(j\omega)|$, $|Y_2(j\omega)|$, $\alpha(\omega)$ and $\beta(\omega)$ as constant over a small frequency range,

$$\frac{d|Y_{cp}(j\omega)|}{d\omega T} = \frac{-|Y_1(j\omega)|[2\alpha(\omega)\sin\omega T - 2\beta(\omega)\cos\omega T]}{2\{1 + |Y_1(j\omega)|^2|Y_2(j\omega)|^2 - 2[\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T]\}^{3/2}}$$

$$= 0 \text{ when } \tan\omega T = \frac{\beta(\omega)}{\alpha(\omega)}$$

i.e. $$\omega T = \arctan\frac{\beta(\omega)}{\alpha(\omega)} + k\pi,\ k = 0, 1, 2, 3 \ldots$$

It follows that at such maxima and minima

$$\sin\omega T = \frac{\pm\beta(\omega)}{\sqrt{[\alpha^2(\omega) + \beta^2(\omega)]}} \text{ and } \cos\omega T = \frac{\pm\alpha(\omega)}{\sqrt{[\alpha^2(\omega) + \beta^2(\omega)]}}$$

Thus

$$\alpha(\omega)\cos\omega T + \beta(\omega)\sin\omega T = \frac{\pm\alpha^2(\omega) \pm \beta^2(\omega)}{\sqrt{[\alpha^2(\omega) + \beta^2(\omega)]}} = \pm\sqrt{[\alpha^2(\omega) + \beta^2(\omega)]} = \pm|Y_1(j\omega)||Y_2(j\omega)|$$

Substituting in eqn. (51) gives, for maxima and minima,

$$\left|Y_{cp}(j\omega)\right|_{\substack{max\\min}} = \frac{|Y_1(j\omega)|}{1 \mp |Y_1(j\omega)||Y_2(j\omega)|}$$

These maxima and minima will occur at values of ωT given by

$$\omega T = \arctan\frac{\beta(\omega)}{\alpha(\omega)} + k\pi = \arctan\frac{A(\omega)D(\omega) + B(\omega)C(\omega)}{A(\omega)C(\omega) - B(\omega)D(\omega)} + k\pi$$

i.e. $$\omega T = \arctan\frac{B(\omega)}{A(\omega)} + \arctan\frac{D(\omega)}{C(\omega)} + k\pi$$

= (excess phase shift in delay unit) + (phase shift in feedback unit) + $k\pi$

(8.2) Response of Type-C Integrator to Input Pulse Train

Consideration of eqn. (35) and Fig. 8, which illustrates the mechanism of the formation of the output pulse train, shows that the transform of the Nth output pulse in the output sequence of a practical type-C integrating filter fed with the train of input pulses discussed in Section 3.1.2 is given by

Transform of Nth output pulse

$$= \bar{F}Y_1(p)\varepsilon^{-NpT}[1 + Y_1(p)Y_2(p) + \ldots + Y_1^{N-1}(p)Y_2^{N-1}(p)]$$

$$= \bar{F}Y_1(p)\varepsilon^{-NpT}\left[\frac{1 - Y_1^N(p)Y_2^N(p)}{1 - Y_1(p)Y_2(p)}\right]$$

At this point in the output sequence, the input pulse train ceases and the output begins to decrease as the stored signals die away in the loop of net gain less than unity. Thereafter,

Transform of the $(N + R)$th output pulse

$$= \bar{F}Y_1(p)\left[\frac{1 - Y_1^N(p)Y_2^N(p)}{1 - Y_1(p)Y_2(p)}\right]Y_1^R(p)Y_2^R(p)\varepsilon^{-(N+R)pT}$$

Thus the output pulse sequence is given by

$$S_o(t) = \mathscr{L}^{-1}\frac{\bar{F}Y_1(p)}{[1 - Y_1(p)Y_2(p)]}$$

$$\left\{\sum_{n=1}^{n=N}[1 - Y_1^n(p)Y_2^n(p)]\varepsilon^{-npT} + \sum_{r=1}^{r=\infty}[1 - Y_1^N(p)Y_2^N(p)]Y_1^r(p)Y_2^r(p)\varepsilon^{-(N+r)pT}\right\}$$

(8.3) Signal/Noise Ratio Improvement of Group Filter with Loop Gain of Unity

As shown in eqn. (44), the signal/noise ratio improvement of a group type of integrating filter with ideal components is given by

$$I_{SN} = \sqrt{\left[\frac{(1 + \beta)(1 - \beta^N)}{(1 - \beta)(1 + \beta^N)}\right]} \quad . \quad . \quad . \quad (44)$$

The expression in the form given becomes indeterminate when $\beta = 1$, in which case the limit may be found by rewriting eqn. (44) as

$$I_{SN} = \sqrt{\left[\frac{(1 + \beta)(1 - \beta)(1 + \beta + \beta^2 + \ldots + \beta^{N-1})}{(1 - \beta)(1 + \beta^N)}\right]}$$

Thus I_{SN} for $\beta = 1$ will be

$$\mathscr{L}_{t\,\beta\to 1}\sqrt{\left[\frac{(1 + \beta)(1 + \beta + \beta^2 + \ldots + \beta^{N-1})}{(1 + \beta^N)}\right]} = \sqrt{N}$$

(8.4) Permissible Frequency Shift of Highest Response Peak of Interest

The parameter b of Section 4.1 may be calculated from the amplitude/frequency response characteristics obtained in Section 2. Considering the case of an ideal type-C integrating filter, the amplitude/frequency response characteristic is given by

$$|Y_c(j\omega)| = \frac{1}{\sqrt{(1 + \beta^2 - 2\beta\cos\omega T)}}$$

The maximum value of the response peak is $\frac{1}{1 - \beta}$ and thus, at the 3 dB points,

$$\frac{1}{\sqrt{2}(1 - \beta)} = \frac{1}{\sqrt{(1 + \beta^2 - 2\beta\cos 2\pi bT)}}$$

solution of which gives

$$b = \frac{I}{2\pi T}\text{arc cos} - \left(\frac{1 + \beta^2 - 4\beta}{2\beta}\right) \simeq \frac{1 - \beta}{2\pi T} \text{ when } 1 - \beta \ll 1$$

A similar procedure may be adopted for any of the other frequency responses obtained.

621.391.821: 621.396: 621.317.34

The Institution of Electrical Engineers
Paper No. 3115 E
Jan. 1960

THE MEASUREMENT OF ATMOSPHERIC RADIO NOISE BY AN AURAL COMPARISON METHOD IN THE RANGE 15–500 kc/s

By J. HARWOOD, M.A., Ph.D., and B. N. HARDEN, M.Sc., Associate Member.

(*The paper was first received* 20*th April, and in revised form* 9*th July*, 1959.)

SUMMARY

Atmospheric noise received on a vertical aerial is compared with a locally-generated keyed signal to estimate the level at which the signal is 95% intelligible. This method was used in earlier high-frequency equipment and has been adapted for the range 15–500 kc/s.

The apparatus and its method of operation are described. Results obtained at a number of sites are discussed in relation to more objective measurements. In particular, deduced noise powers are compared with existing world-wide predictions, revealing some differences.

(1) INTRODUCTION

The comparison of early measurements of atmospheric radio noise levels at various locations was often made difficult by the diversity of characteristics which were chosen to represent the level. In order to overcome this difficulty, an equipment was designed by H. A. Thomas[1] and distributed in 1945 to a number of sites throughout the world, so that the same noise parameter would be measured at all the sites. A parameter was chosen which could be measured with easily obtained commercial equipment, and which had direct value in the assessment of the interference caused by the noise to Morse communication, namely, the level of a low-speed Morse signal which could be read with 95% intelligibility through the incoming noise. The measurements were made hourly in the frequency range 2–20 Mc/s by comparing the atmospheric noise received on a vertical aerial with the strength of a locally generated, keyed signal.

Similar apparatus was developed later for operation at low frequencies (15–500 kc/s), and was installed during 1954 and 1955 at a number of sites in various parts of the world. The equipment is described, and results obtained with it are related to other noise measurements and compared with predictions issued in a report[2] of the International Radio Consultative Committee.

The report gives expected values of a specific parameter of the atmospheric noise for all parts of the world, at all times of the day and night and all seasons of the year, for frequencies from 10 kc/s to about 30 Mc/s. The predicted values in the low-frequency region were based principally on measurements taken at a few places, e.g. in England, the United States, South Africa and the Arctic, values for other locations being deduced by interpolation using a knowledge of thunderstorm distribution and radio-propagation characteristics. References to the sources of the data and the use of predicted values in relation to various types of service are given in the report.

(2) DESCRIPTION OF EQUIPMENT

The main units are a vertical mast aerial, a receiver, and a signal generator: the last two units are suitable for mounting in a 19 in rack, and have panels of height $8\frac{3}{4}$ and $11\frac{1}{2}$ in respectively. Atmospheric noise received on the aerial is applied to the receiver, together with keyed signals from the signal generator adjacent to the receiver. The signals are made audible as a 1 kc/s tone in the headphones, using a beat oscillator in the receiver. Adjustment of signal intensity relative to that of the noise is made by means of a stepped attenuator in the signal generator, which covers amplitudes from below the internal receiver noise level to levels in excess of any atmospheric noise likely to be encountered.

Keying is effected by a simple electronic circuit in the signal generator, so that, in comparison with the earlier high-frequency equipment, the keying arrangements are considerably simplified. The signal, consisting of a series of Morse dashes (two per second), is adjusted in level until the operator estimates that a true Morse signal of the same amplitude would be just intelligible.

A block schematic of the equipment is shown in Fig. 1. Being designed specifically for low-frequency atmospheric noise measurement, the equipment involves a few unconventional features which are briefly described in the following Sections.

(2.1) Receiver

A superheterodyne receiver is used with four ranges of frequency covering the band 15–500 kc/s. By means of two crystal-tuned circuits at the intermediate frequency of 1·6 Mc/s, a bandwidth of 300 c/s is obtained (between 3 dB points of the response curve). The r.f. section comprises an amplifier and a cathode-follower, the grid circuit of each being tuned. The mixer is a low-impedance diode bridge to which the local oscillator and r.f. signals are coupled through a toroidal dust-core transformer. Provision of a balanced mixer was necessary to reject voltages at the local-oscillator frequency, which would otherwise be passed through the i.f. section at spurious crystal resonances near the intermediate frequency.

The i.f. section consists of three tuned-amplifier stages, two of which incorporate matched crystals, and a cathode-follower output which supplies the i.f. signals at low impedance to two diode detectors. A beat-oscillator signal can be injected at the cathode-follower stage. One of the diodes is used in a conventional a.f. detection circuit connected via an amplifier to the headphones, while the other is d.c. coupled to an external socket to permit measurement of parameters of the noise envelope such as the average or r.m.s. voltage. The main control of gain is by variation of bias at the r.f. and one i.f. amplifier. There is a subsidiary preset l.f. gain control.

(2.2) Signal Generator

The main sections of the signal generator comprise an oscillator in an inner screened box, a circuit providing keying and modulation facilities, and a stepped attenuator. A separate 1·6 Mc/s sweep-frequency generator is provided in the same unit for receiver alignment.

A 2-valve oscillator covers the frequency range 15–500 kc/s in four ranges corresponding to those of the receiver, and a thermistor is used for amplitude stabilization. By means of a 'set-carrier' potentiometer, the output is adjusted over a small range, and can be modulated or keyed electronically by a diode

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

The paper is an official communication from the Radio Research Station, Department of Scientific and Industrial Research.

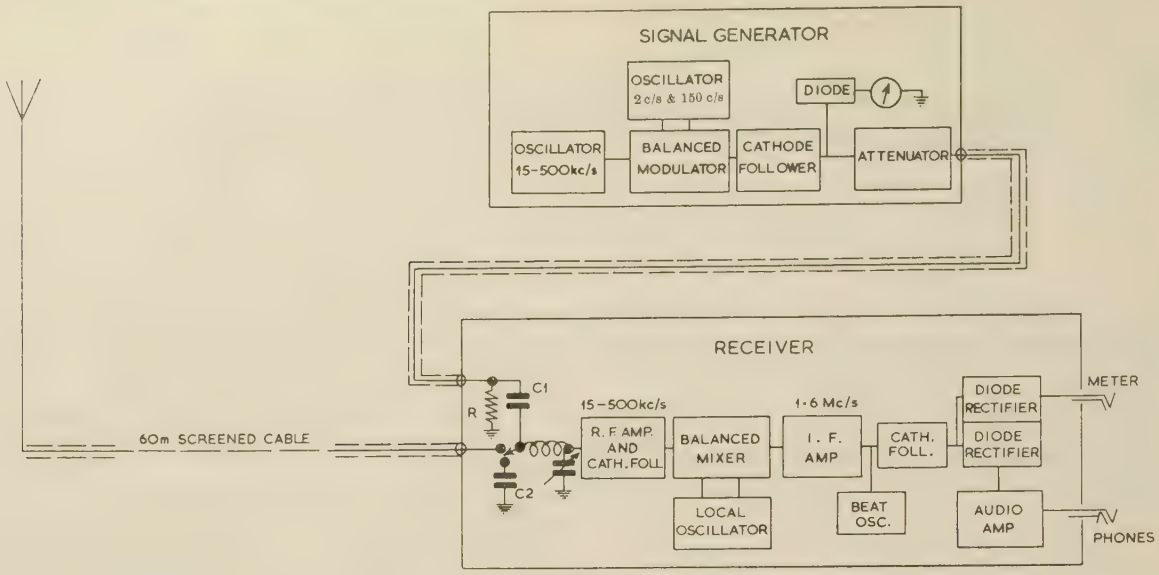


Fig. 1.—General arrangement of equipment.

R = Signal-generator matching resistor (40 ohms).
C_1 = Signal-generator coupling capacitor (value equal to aerial capacitance to earth).
C_2 = Dummy aerial capacitor.

bridge circuit and phase-shift oscillator (Fig. 2). The signal is then applied to a cathode-follower, across whose output are connected the stepped attenuator and a 'set-carrier' meter circuit. A pair of 6-position units in series act as fine and coarse attenuation controls, providing 36 steps each of 3 dB, and the attenuator is terminated by a 40-ohm resistor in the receiver; the maximum output is 10 mV.

Three types of output are used, namely

(*a*) Continuous-wave signal (no modulation).
(*b*) Signal amplitude modulated at 150 c/s for tuning to receiver.
(*c*) Signal switched at 2 c/s to represent a Morse signal.

The modulated and keyed outputs are obtained by operating the phase-shift oscillator at 150 or 2 c/s, respectively. In the latter case, a large 2 c/s sinusoidal voltage is applied to the bridge so that approximately square-wave modulation results; this technique was chosen in preference to the use of a relay or multivibrator because of the audible 'key-click' effect which was often present when accurately square-wave switches were used. The continuous-wave signal is obtained by applying fixed bias to the diode bridge.

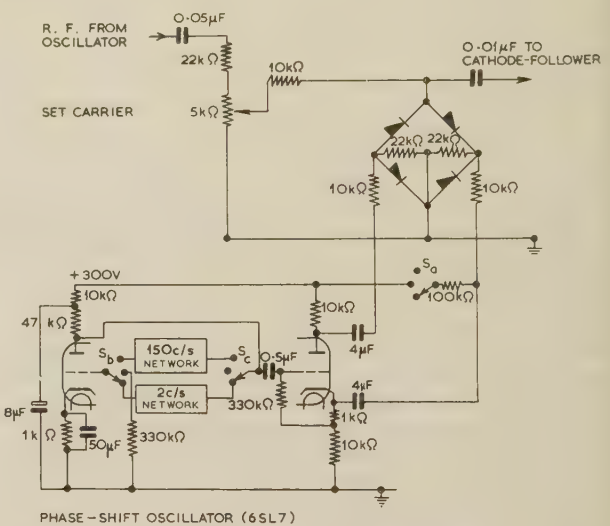


Fig. 2.—Signal-generator keying circuits.

(2.3) Aerial and Receiver Input Arrangements

Calibration of the equipment, or the conversion of signal-generator outputs to equivalent field strengths at the aerial, depends on the aerial and circuits prior to the first stage of the receiver; these are shown in Fig. 1. The aerial is a vertical mast whose height (about 13 m) is less than one-fiftieth of a wavelength at the highest frequency used, and its impedance is effectively a pure capacitance. The base of the aerial is directly connected to the inner conductor of the concentric cable (60 m long), which also behaves as a capacitance and is used as part of a series-tuned circuit at the input to the receiver. At the lowest frequencies it was found desirable to damp the input tuned circuit, so that its bandwidth would not be comparable with that of the i.f. section. Although the gain is thereby reduced, the sensitivity is adequate for recording the relatively high noise levels at these frequencies.

The signal-generator output is coupled to the receiver input circuit through a capacitor C_1 approximately equal in value to the aerial capacitance. Thus the conversion of signal-generator output to equivalent field strength depends only on the effective height of the aerial and on any difference between C_1 and the aerial capacitance.

To assist in tuning the signal generator to the receiver, a switch is provided to replace the aerial cable by a dummy load without affecting the first-stage tuning.

(3) OPERATION OF EQUIPMENT

The procedure in making a noise measurement is briefly as follows:

(*a*) The receiver is tuned to a frequency, near to the desired frequency, at which there is negligible station interference, so that only atmospheric noise is being received.
(*b*) The signal generator is tuned to the receiver, using 150 c/s modulation.
(*c*) The keying is switched on, together with the receiver beat oscillator, and the stepped attenuator is adjusted until it is estimated that about 95% of the keyed signal would be intelligible if it were true Morse.

Measurements are made, in general, at each hour of the day on five frequencies. The frequencies were chosen to cover most of the range 15–500 kc/s, and to be located where least station interference was expected; the values finally chosen were 18, 30, 135, 220 and 400 kc/s.

The accuracy of measurement depends on the type of noise encountered; for example, thermal-type noise can be assessed more accurately than the impulsive type of noise found at the lower frequencies. Even in the latter case, however, a trained observer can be expected to obtain an accuracy of ±3 dB in an individual measurement.

(4) RESULTS

(4.1) Stations in Operation

The equipment has been distributed to eight observing stations which are listed in Table 1.

Table 1

OBSERVING STATIONS

Station	Latitude	Longitude
Accra, Ghana	05° 34′ N	00° 14′ W
Colombo, Ceylon	07° 13′ N	79° 52′ E
Fanning Island, Pacific Ocean	03° 55′ N	159° 23′ W
Halley Bay, Antarctica ..	75° 30·5′ S	26° 36′ W
Johannesburg, South Africa	26° 06′ S	27° 55′ E
Nicosia, Cyprus	35° 10′ N	33° 23′ E
Port Stanley, Falkland Is. ..	51° 45′ S	57° 58′ W
Tatsfield, England	51° 17′ N	00° 00′

In addition, equipment has been supplied to the Danish authorities for operation in Greenland.

Hourly observations are sent to the Radio Research Station at Slough for analysis, except from a few stations where data are first summarized for each month.

(4.2) Method of Analysis

For each station, noise measurements at each of the five frequencies have been analysed as follows. Upper-decile, median and lower-decile values were deduced from readings at a particular hour over specific periods. Until May, 1957, these values were obtained for each hour of the day over one-month periods and then averaged over seasons of three months. Later, median and decile values were deduced for 2-hour intervals over the seasonal periods. All data were finally plotted to show diurnal variations for each frequency and season.

The results at this stage give minimum levels of low-speed Morse transmission intelligible through the noise received on an omnidirectional aerial. To deduce the interference caused to other types of transmission, and to compare values with those measured by other methods, it is necessary to relate the assessed levels to more objective parameters. This has been done by comparing the results from Tatsfield with observations from Slough of various parameters of the noise at frequencies between 10 and 135 kc/s, taken within the same hour. The Slough records were taken at intervals between 1953 and 1955[3] and during the International Geophysical Year.

A statistical evaluation of results over periods of three months indicates reasonable agreement in the shape of the diurnal variations of the noise levels at Tatsfield and of the average voltage of the noise envelope measured in a bandwidth of 300 c/s at Slough. In general, for all frequencies and time blocks, the minimum intelligible signal level measured in a bandwidth of 300 c/s is found to be about 6 dB above the peak value of the c.w. field equivalent to the average of the noise envelope.

Further conversion factors based on experimental work at Slough enable the results to be expressed in terms of the r.m.s. value of the noise field strength or, for more convenient comparison with other measurements and with predictions, in terms of noise power.

The C.C.I.R. report[2] gives predicted levels in terms of a basic parameter F_a, which may be considered as an effective aerial noise figure and is related to the noise power available from a short, vertical, earthed, loss-free aerial. The available power is independent of the length of an aerial which is short compared with the wavelength.

If the large, rapid fluctuations of noise power are averaged for several minutes, the average values are, in general, found to be nearly constant during any given hour. F_a is the median value of this average power during a period of one hour and is expressed in decibels relative to the thermal noise power available from the aerial if it were at a specified temperature, T, i.e. kTB watts, where k is Boltzmann's constant ($=1{\cdot}38 \times 10^{-23}$ joules per deg K), B is the effective bandwidth in cycles per second, and T is the reference temperature (chosen to be 288° K in the C.C.I.R. report). F_a is independent of bandwidth, since the available noise power and the reference level are both proportional to bandwidth, and F_a, in decibels, is related to the r.m.s. field strength at the aerial by the equation[2, 4]

$$F_a = E_n + 65{\cdot}5 - 20 \log_{10} f$$

where E_n is the r.m.s. field strength for a 1 kc/s bandwidth in decibels above 1 μV/m, and f is the frequency in megacycles per second. All experimental results have therefore been expressed in terms of F_a, using the assumption, derived from experimental factors, that F_a is the measured minimum field intensity in decibels above 1 μV/m for 95% intelligibility of slow Morse in a 300 c/s bandwidth + C, where C is given in Table 2.

Table 2

CONVERSION FACTOR, C

Frequency, kc/s	C, dB
18	108
30	101
135	84
220	81
400	75

Within the limits set by the accuracy of the original observations no significant trends in the variation of C with time of day or season have been revealed, and the conversion factor has been assumed to be dependent only on frequency.

The r.m.s. value of the noise field strength in a bandwidth of 1 kc/s can be deduced from the above relation.

(4.3) Typical Results

Fig. 3 shows values for three stations giving results typical of different areas of the world. Median levels, plotted as a function of time of day, frequency and season, are superposed for Fanning Island, Johannesburg and Tatsfield using data taken up to the end of May, 1957, over periods of 11, 22 and 29 months, respectively. For convenience in comparison these are grouped for seasonal periods obtained, as in the C.C.I.R. report,[2] by dividing the year into four seasons of three months as shown in Table 3.

The greatest diurnal changes occur at all frequencies and seasons at Johannesburg, which is close to the major thunderstorm centres in Africa. At Fanning Island which is near the equator

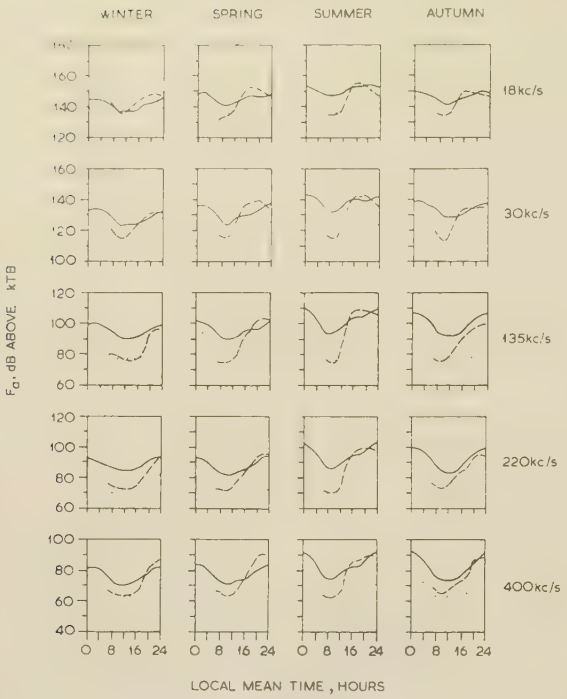


Fig. 3.—Median values of F_a.
····· Fanning Island.
– – – – Johannesburg.
—— Tatsfield.

Table 3

Months	Seasons	
	Northern Hemisphere	Southern Hemisphere
December, January, February..	Winter	Summer
March, April, May	Spring	Autumn
June, July, August	Summer	Winter
September, October, November	Autumn	Spring

and surrounded by large ocean areas far from the major storm centres, both diurnal and seasonal changes are small. In general, the minimum noise levels at Johannesburg approximate to the minimum levels registered at Fanning Island, but the peak levels (except in winter) are appreciably greater. Although no measurements are taken at Johannesburg from 2300 to 0700 hours, the maximum noise levels in summer are apparently reached during the afternoon, and this trend continues at the lower frequencies in spring and autumn as well. At Fanning Island the peak values occur during the early hours of the morning at all seasons.

The Tatsfield results show peak values of the same order as those obtained at Johannesburg, but the minima do not fall to the levels recorded at both Fanning Island and Johannesburg. Comparison of Tatsfield values with measurements at Slough indicates that the former are too high from about 0600 to 1500 hours, probably because of interference from unwanted stations and, to a smaller extent, man-made noise. It is believed that the Slough results at 135 kc/s are themselves contaminated by signal interference at the lowest levels but not nearly to the same degree as the Tatsfield records. Consequently the true diurnal variation at Tatsfield would seem to be greater than the results indicate. Allowance for this effect was made in deducing the conversion factor, C, of Table 2. Upper-decile values at Tatsfield show the afternoon rise rather more than do the medians, particularly during the summer.

(4.4) Relationship to High-Frequency Measurements and to Predictions

Combination of the present low-frequency measurements with earlier measurements at high frequencies[5,6] permits a reasonable coverage to be made of the radio-frequency spectrum between 10 kc/s and 20 Mc/s. Comparisons have been made with the C.C.I.R. predictions of noise fields. In the C.C.I.R. report,[2] estimated values of F_a and E_n are presented for 24 time blocks, each consisting of a particular four-hour period during the day (0000–0400, 0400–0800 hours, etc.) throughout a season. For each time block every part of the world is allotted a contour number which corresponds to the estimated time-block median value of F_a in decibels at a frequency of 1 Mc/s. Predicted noise values at other frequencies and upper- and lower-decile values can then be derived from subsidiary curves.

An example is given in Fig. 4 illustrating both the spectrum coverage and the comparison with predictions. Experimental values of F_a are those obtained at Johannesburg during the period 2000–2400 hours, between August, 1955, and May, 1957, at frequencies of 18, 30, 135, 220 and 400 kc/s using the low-frequency equipment, and between December, 1951, and November, 1956, at frequencies of 2·5, 5, 10 and 15 Mc/s using the high-frequency equipment. Median values for each season are shown superposed on the curves of predicted values of F_a plotted as a function of frequency and contour number. Deviations from predictions occur, and this evidence would have to be considered in conjunction with other data to decide whether any major change in the curves or contours was necessary.

(5) SUMMARIZED RESULTS

Low-frequency noise values for five stations (Colombo, Fanning Island, Johannesburg, Nicosia and Tatsfield) recorded up to the end of May, 1957, have been summarized to give a direct comparison with C.C.I.R. predictions. For any time block and frequency, measured and estimated values may differ in two respects: departure from the estimated frequency dependence of F_a with contour number and error in the estimated contour. Plotting of experimental values of F_a for each time block and station as a function of frequency (as in Fig. 4) first revealed that noise levels were rarely observed below the values of man-made noise at a quiet receiving site predicted in the C.C.I.R. report (shown in Fig. 4), and that when sufficiently low levels were recorded, particularly in the time blocks from 0800 to 1200 hours, the frequency dependence above about 100 kc/s was that shown by this curve.

With due allowance for this effect, a straight line giving the best fit to the plotted experimental data was then drawn for each station and time block. The differences in the values of F_a at 10 and 100 kc/s on these lines were then plotted as a function of the value of F_a at 100 kc/s and compared with values similarly taken from the C.C.I.R. curves. A convenient representation was thus given of any experimental departure from the slopes of the C.C.I.R. curves in the low-frequency region where the variations of F_a with frequency are approximately linear for all contour numbers.

The C.C.I.R. report gives two sets of curves, one for times from 0400 to 2000 hours in spring and summer and from 0800 to 1600 hours in autumn and winter (periods of daylight), and the other for remaining darkness periods. The experimental data show no significant difference between the two time groupings and can be adequately represented by the solid line in Fig. 5, the standard deviation being ± 3 dB. Also given in Fig. 5 are

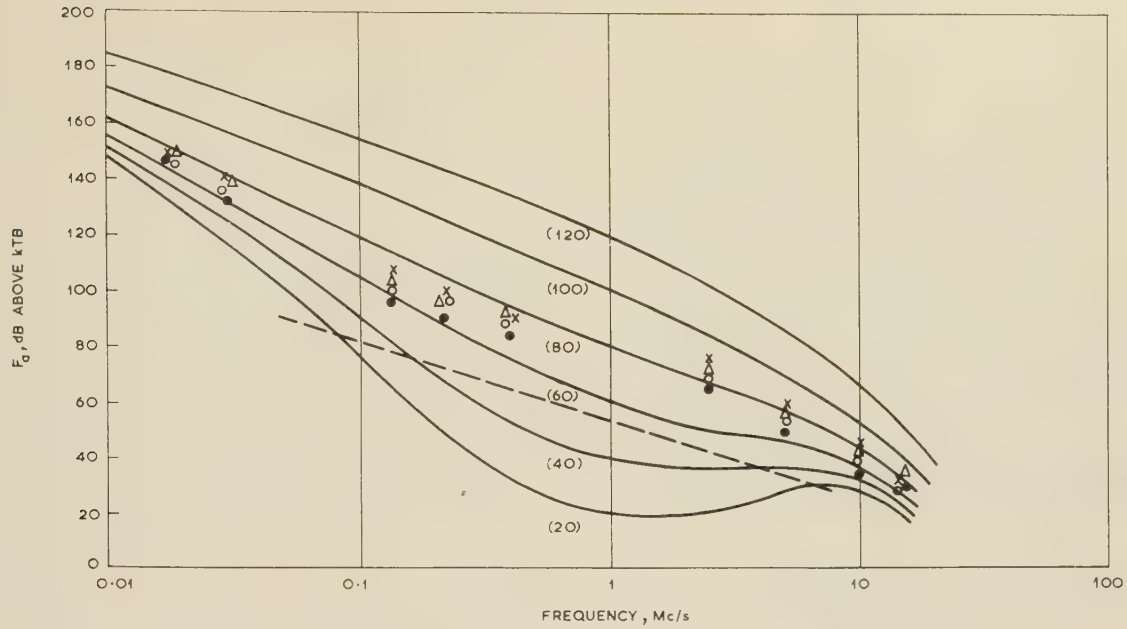


Fig. 4.—Comparison of experimental and predicted median values of F_a at Johannesburg (2000–2400 hours).

Full lines show predicted values as a function of frequency and contour number. (Contour number corresponds to the estimated time-block median value of F_a at 1 Mc/s.)
Dotted line shows expected values of man-made noise at a quiet receiving site.

Period	Experimental Values	Predicted contour numbers
Dec. to Feb. ..	×	85
Mar. to May ..	○	78
June to Aug. ..	●	70
Sept. to Nov. ..	△	78

the appropriate C.C.I.R. curves for daylight and darkness. Over the range of values of F_a at 100 kc/s between 80 to 120, the frequency dependence of the experimental results tends to conform to that of the day-time set of curves for higher values of F_a and to the night-time set for lower values of F_a.

Table 4 gives, for each station and time block, a value of F_a at 100 kc/s deduced as above and the departure from the estimated C.C.I.R. value. At Colombo the differences are generally small except in the afternoons and evenings between December and February and from June to August. At Fanning Island the values are appreciably below the predictions at all times. Little variation in diurnal changes occurs throughout the year. Results from Johannesburg reveal that predicted values are, in general, too high, whereas those from Nicosia indicate an opposite tendency. At Tatsfield there is general agreement, the principal differences occurring in the early afternoon except in winter, and during the late morning in summer and autumn.

Table 5 lists the ratio, D_u, in decibels of the upper-decile to the median value and the ratio, D_l, of the median to the lower-decile value for each time block. These have been grouped as shown for comparison with the C.C.I.R. predictions in which D_u and D_l as a function of frequency are given by two curves, one applying to D_l for all time blocks and to D_u for the time blocks 0000–0400 and 2000–2400 hours, and the other applying to D_u for the time blocks between 0400 and 2000 hours. Experimental values of D_u and D_l show some scatter from the C.C.I.R. curves, particularly for D_u at the higher frequencies for time blocks between 0400 and 2000 hours. The general tendency is for the experimental ratios to be falling with increase of frequency above 135 kc/s instead of still slowly rising as predicted.

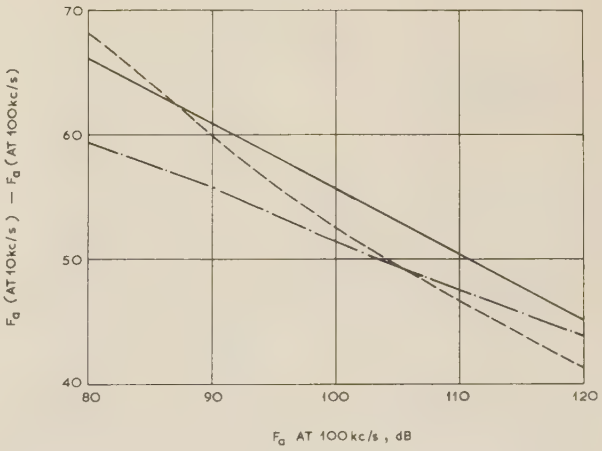


Fig. 5.—Relation between the frequency dependence of F_a and its absolute value at 100 kc/s.

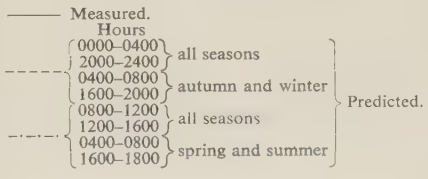


Table 4

COMPARISON BETWEEN MEASURED AND PREDICTED MEDIAN VALUES OF F_a AT 100 KC/S

Station and observation period	Season	F_a at 100 kc/s, dB											
		0000–0400 hours		0400–0800 hours		0800–1200 hours		1200–1600 hours		1600–2000 hours		2000–2400 hours	
		M	D	M	D	M	D	M	D	M	D	M	D
Colombo, January, 1956, to May, 1957	Dec.–Feb.	110	−5	106	−3	100	4	98	−7	104	−5	108	−7
	Mar.–May	120	3	112	−4	103	4	103	−1	115	−5	122	5
	June–Aug.	116	2	108	−6	96	1	99	−11	106	−12	117	2
	Sept.–Nov.	115	0	106	1	98	1	100	−2	111	2	117	2
Fanning Island, July, 1956, to May, 1957	Dec.–Feb.	102	−9	99	−9	88	−8	86	−17	89	−21	95	−16
	Mar.–May	98	−13	96	−13	89	−8	88	−10	89	−21	94	−17
	June–Aug.	101	−11	99	−9	86	−6	86	−18	86	−24	94	−18
	Sept.–Nov.	102	−9	98	−11	87	−9	86	−11	86	−24	96	−15
Johannesburg, August, 1955, to May, 1957	Dec.–Feb.	No observations				85	−22	109	−8	112	−16	114	−11
	Mar.–May					85	−16	98	−6	104	−9	108	−11
	June–Aug.					86	−6	86	−12	93	−12	106	−7
	Sept.–Nov.					87	−14	97	−7	106	−13	112	−7
Nicosia, May, 1956, to May, 1957	Dec.–Feb.	118	5	110	4	104	10	104	5	112	7	118	5
	Mar.–May	118	5	111	−1	105	10	106	8	110	−3	116	3
	June–Aug.	114	1	109	4	106	14	105	0	110	−6	114	1
	Sept.–Nov.	114	5	107	5	100	5	98	−1	107	2	113	4
Tatsfield, January, 1955, to May, 1957	Dec.–Feb.	108	−1	105	0	99	1	98	0	101	−4	107	−2
	Mar.–May	110	−3	103	−3	97	2	100	5	105	−3	110	−3
	June–Aug.	116	3	106	2	105	13	110	9	113	3	117	4
	Sept.–Nov.	114	1	110	5	102	7	102	4	108	−2	114	1

M: Measured value.
D: Difference between measured and predicted value.

Table 5

COMPARISON OF MEASURED AND PREDICTED VALUES OF D_u AND D_l

	Frequency	Predicted values	Measured values				
			Colombo	Fanning Island	Johannesburg	Nicosia	Tatsfield
	kc/s						
D_u, in dB, for time blocks 0000–0400 h 2000–2400 h	18	5	4	6	6	8	9
	30	6	6	7	8	7	10
	135	10	10	10	12	8	12
	220	11	7	10	11	8	11
	400	12	8	11	11	9	11
D_u, in dB, for time blocks 0400–0800 h 0800–1200 h 1200–1600 h 1600–2000 h	18	6	6	6	8	7	10
	30	7	8	8	10	8	12
	135	16	10	12	15	8	15
	220	17	8	7	12	10	13
	400	18	9	7	13	9	12
D_l, in dB, for all time blocks	18	5	7	4	7	9	10
	30	6	8	7	8	11	10
	135	10	7	9	9	10	11
	220	11	7	6	9	10	9
	400	12	8	4	8	10	8

(6) CONCLUSIONS

A description has been given of a compact low-frequency noise-measuring equipment which uses the same basic technique as has previously been applied at high frequencies.[1] The equipment is in use at a number of sites, and the measurements show directly the degree of interference which would be caused to Morse communication. It is possible to relate the data to more objective parameters of the noise, and, in particular, to compare deduced values of the noise power with existing predictions of the same parameter.

Experimental results at frequencies from 18 to 400 kc/s have been summarized to show the changes in noise power with time at five observing stations. Differences at 100 kc/s between measured median values and those predicted by the C.C.I.R. have been indicated. The dependence of the median noise level on frequency (frequency law) has been investigated. The results

indicate that the use in the predictions of different frequency laws to distinguish between periods of daylight and darkness is not justified at present. Ratios of upper and lower decile to median values show smaller variations with frequency than have been estimated. In general, the lower limit of the measured noise is about the level predicted for man-made noise at a quiet receiving location.

(7) ACKNOWLEDGMENTS

Many of the measurements have been carried out by the following organizations: the British Broadcasting Corporation; Cable and Wireless, Ltd.; Overseas Telecommunciation Service, Ceylon; the South African Broadcasting Corporation. Their assistance is much appreciated.

The advice of Mr. F. Horner who supervised the work is gratefully acknowledged.

The work described in the paper was carried out as part of the programme of the Radio Research Board and is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

(8) REFERENCES

(1) THOMAS, H. A.: 'A Subjective Method of Measuring Radio Noise', *Proceedings I.E.E.*, Paper No. 1006 R, September, 1950 (**97**, Part III, p. 329).

(2) C.C.I.R. Report No. 65: 'Revision of Atmospheric Radio Noise Data' (International Telecommunication Union, Geneva, 1957).

(3) HARWOOD, J.: 'Atmospheric Radio Noise at Frequencies between 10 kc/s and 30 kc/s', *Proceedings I.E.E.*, Paper No. 2619 R, May, 1958 (**105** B, p. 293).

(4) CRICHLOW, W. Q.: 'Noise Investigation at V.L.F. by the National Bureau of Standards', *Proceedings of the Institute of Radio Engineers*, 1957, **45**, p. 778.

(5) THOMAS, H. A.: 'Some Measurements of Atmospheric Noise at High Frequencies', *Proceedings I.E.E.*, Paper No. 1007 R, September, 1950 (**97**, Part III, p. 335).

(6) 'Measurements of Atmospheric Noise at High Frequencies', Radio Research Special Report No. 26 (H.M. Stationery Office, 1953).

621.385.624.2

The Institution of Electrical Engineers
Paper No. 3130 E
Jan. 1960

©

AN EXPERIMENTAL INVESTIGATION OF A TWO-CAVITY KLYSTRON OPERATING UNDER LARGE-SIGNAL CONDITIONS

By I. M. STEPHENSON, Ph.D., M.Sc., B.Sc.

(*The paper was first received* 31*st March, and in revised form* 25*th June*, 1959.)

SUMMARY

An experimental 2-cavity klystron with a variable drift space has been constructed to investigate velocity-modulated electron beams. The operating frequency is 300 Mc/s. Previous experimental work with this valve has shown that the space-charge-wave theory accurately predicts both the optimum drift length and the output r.f. current, provided that small signals are used. Recent theoretical work on large-signal conditions has enabled some experimental results with large signals to be compared with calculations for similar, though not quite identical, conditions. Both the experiments and the theory show that the optimum drift length is considerably reduced when the input signal is large. The parameter that determines whether the signal is large is not merely the ratio of the modulating r.f. voltage to the beam accelerating voltage but is this quantity multiplied by the ratio of signal frequency to plasma frequency.

LIST OF SYMBOLS

a = Drift tube radius.
b = Beam radius.
I_0 = Direct component of beam current.
I_i = Alternating component of beam current.
r = A radius.
R_1 = Normalized drift tube radius = $\gamma_0 a$.
R = Normalized radius = $\gamma_0 r$.
T = Space-charge-wave coefficient.
V_0 = Beam accelerating voltage.
V_1 = Gap alternating voltage.
v_0 = Velocity of d.c. beam.
z = Distance measured along beam axis.
α = Modulation coefficient, V_1/V_0.
β = A gap factor.
β_r = Gap coupling coefficient at a radius r.
$\gamma_0 = \omega/v_0$.
ρ_0 = Charge density of unmodulated electron beam.
$\omega = 2\pi f$.
ω_p = Plasma angular frequency = $\sqrt{(e\rho_0/m\epsilon_0)}$.

$$\omega_q = \text{Modified plasma frequency} = \frac{\omega_p}{\sqrt{\left[1 + \left(\frac{2}{\gamma_0 b}\right)^2\right]}}.$$

(1) INTRODUCTION

The earliest quantitative theoretical work on klystron behaviour was that of Webster,[1] who devised the well-known ballistic theory. This theory was quite satisfactory for the earlier klystron valves which used relatively low-current-density electron beams where the space-charge forces are small enough to be neglected. More recent advances in the techniques of klystron construction have resulted in the use of high-current-density beams, and under such conditions the Webster theory can give erroneous results. In these beams the bunching process is almost completely controlled by space-charge forces under small-signal conditions, and the space-charge-wave theory of Ramo[2] and Hahn[3] can be used. There is obviously a range of beam-current densities in which the bunching is partly controlled by space charge and partly by ballistic considerations, and for many practical valves this region corresponds to that in which the r.f. voltage applied to modulate the beam is comparable with the beam accelerating voltage.

The problem of a large-signal theory which includes space-charge effects has recently been attacked numerically by Webber,[4] who has produced a number of computed results for a two-cavity klystron operating under large-signal conditions. This analysis shows the important transition from the space-charge-controlled condition to the ballistic condition. During a recent experimental study[5] of a klystron, which was designed to show the space-charge-wave operation, a number of measurements were made using signals which were too large for true space-charge-limited conditions to be valid. In the paper, the experimental results are compared with Webber's calculations and it is seen that the transition from one mode of operation to another is correctly predicted by the theory.

Although a number of simplifying assumptions are made which are not realized in practice, it is found that a reasonable quantitative agreement is obtained between the experimental and theoretical results, particularly as regards the optimum drift length, a factor which is of extreme practical importance in the design of a klystron valve which is to have a high efficiency.

Experiments with variable-drift-length klystrons have also been carried out by Chodorow[6, 7] and his fellow-workers and by Mihran.[8] Their results are discussed in Section 5.

The theoretical work has been recently extended to cover the case of a multi-cavity valve, and although no experimental work has yet been performed to test this extended theory, the reliability of the theory in its original form in predicting the results in the two-cavity case would seem to be a good indication of its usefulness in studying the multi-cavity valve.

(2) EXPERIMENTAL ARRANGEMENTS

The experimental klystron used for the large-signal measurements was that used in the small-signal measurements described previously[5] and will not be described in detail here. The use of 300 Mc/s as the operating frequency simplifies both the problems of construction and of measurement. An indication of the voltages across the buncher and catcher gaps was obtained by two small loops in the cavities which fed microammeters via crystal rectifiers. The relationship between the rectified crystal current and the r.f. voltage was established before the cavities were placed in the vacuum system by using a valve voltmeter.

The electron beam was 2 cm in diameter and the catcher cavity could be moved so as to vary the drift length from a few millimetres to 15 cm.

The gap on the modulating cavity was gridded whilst that on the output cavity was gridless. Experience with a gridded output gap showed that heavy loading of the output cavity was caused by secondary emission from the grid wires. Although this loading could be measured and taken into account, it was

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

Dr. Stephenson is in the Electrical Engineering Department, University College, London.

found to be a very variable effect and so an ungridded gap was used. It was necessary to know the impedance of the output cavity under conditions of beam loading in order to evaluate the gap current from the measured gap voltage.

The relationship between the current appearing at the output gap and the r.f. current in the beam is rather more complex, and this problem is discussed in Section 4.3.

(3) PREVIOUS THEORETICAL TREATMENT

Previous theoretical work on klystron behaviour has been well reviewed and may be roughly divided into four types:

(*a*) Ballistic theory with no space charge.
(*b*) Ballistic theory with space-charge correction.
(*c*) Single and multi-mode space-charge-wave theory.
(*d*) Large-signal ballistic theory with space charge.

To those theories that consider space charge may be added the additional complication of finite beam and drift-tube sizes.

In most modern klystrons the space-charge forces cannot be neglected so that (*a*) is not accurate. (*b*) and (*c*) may be used provided that the limitations imposed by the theory are observed in practice. The signal must be truly 'small', in that the electrons must not overtake each other. Provided that the drift length is sufficiently great the condition for overtaking depends upon the space-charge density in the beam as well as on the ratio V_1/V_0, and under conditions of small space charge, overtaking may occur even when the depth of modulation is small. As Webber points out, when overtaking occurs the sign of the space-charge force reverses and this is not allowed for in (*b*) and (*c*).

The fourth type of theory, (*d*), allows for the effects of electron overtaking but, unlike (*c*), does not allow for radial variation of the r.f. current. However, it gives much useful information as to the effect of the depth of modulation on such important parameters as the optimum drift length and ratio of alternating to direct beam current. There is another assumption, which is that the velocity modulation at the input gap is sinusoidal. In practice a sinusoidal voltage is applied to the input gap. If this voltage is comparable in magnitude with the beam accelerating voltage, the resulting velocity modulation will not be truly sinusoidal.

To include these non-linear effects and those of radial variations would make the computations virtually impossible, but their neglect must be borne in mind when comparing the experimental results with those obtained theoretically.

(4.1) Experimental Results

The purpose of the previous experimental work was to confirm the small-signal space-charge-wave theory, and the modulation depth, α, was never allowed to exceed 0·3. The experimental results showed that when the space-charge density in the beam was very small the optimum drift length became dependent upon the modulation depth, even when the former was small compared with the beam voltage. These results, together with those taken at slightly higher depths of modulation, showed that the important limiting factor for true small-signal operation was $(\omega/\omega_q)\alpha/2$ rather than α and that space-charge-wave theory failed when the former exceeded unity. When α increased so that $(\omega/\omega_q)\alpha/2 > 1$, the optimum drift length decreased, as predicted by the ballistic theory. It was also noted that the variation of r.f. beam current with drift length was in agreement with both the simple ballistic and the space-charge-wave theories for values of drift length which are very much less than the optimum for the first maximum of r.f. current.

To investigate this transition between the two modes of operation, a number of experiments were conducted using values of V_1/V_0 up to unity. Fig. 1 shows how the drift length is affected by the depth of modulation.

Fig. 2 summarizes these curves by showing how the optimum drift length and maximum output voltage vary with the modulation depth. It is seen that after a certain point the output no longer varies linearly with the input. In a practical valve the reducion in the slope of the curve would be modified by the output being fixed in position while the optimum drift length is decreasing. This is illustrated by Fig. 3, where the cavity was fixed in the optimum position for small-signal operation while the modulation was increased. For large signals the increase in the r.f. beam current is more than offset by the reduction in

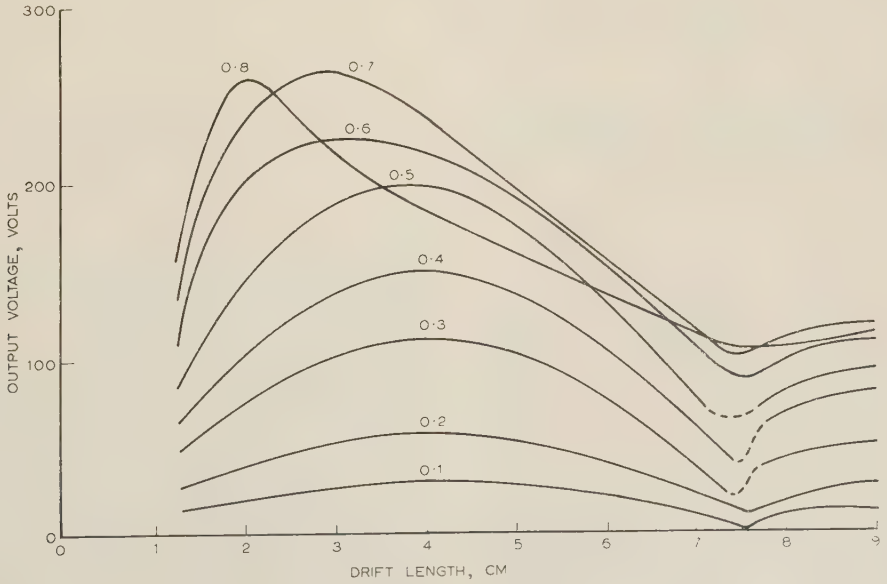


Fig. 1.—Variation of output voltage with drift length.
$V_0 = 200$ volts, $I_0 = 20$ mA
The figures on the curves show values of α.

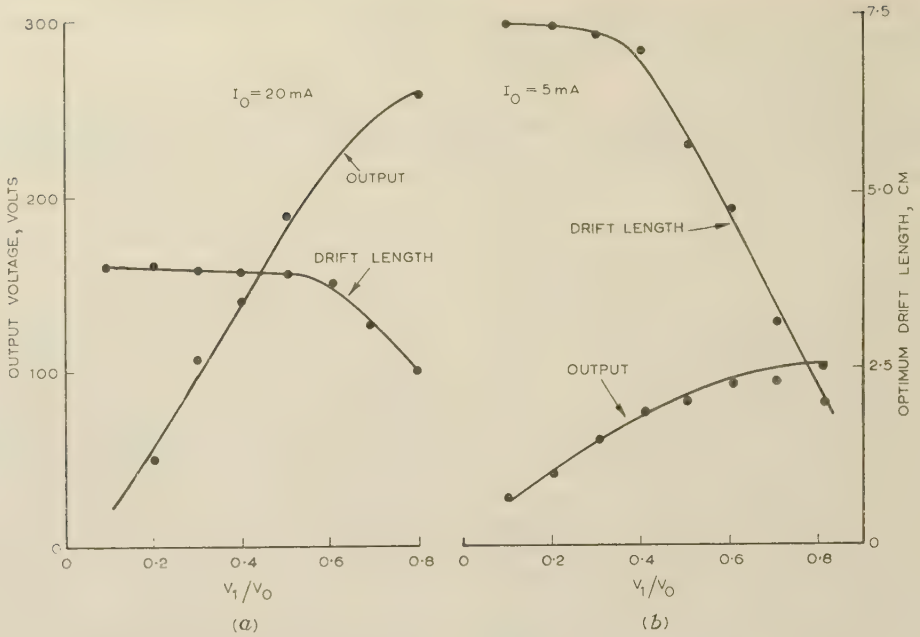


Fig. 2.—Variation of optimum drift length and maximum output voltage with α.
(*a*) $V_0 = 200$ volts, $I_0 = 20$ mA. (*b*) $V_0 = 200$ volts, $I_0 = 5$ mA.

output due to the cavity not being in the optimum position, so that the output from the cavity actually decreases even though the modulation, and hence the maximum r.f. beam current, is being increased.

A number of large-signal results were obtained in addition to those actually shown here, and the results obtained from these measurements are considered in the next Section.

(4.2) Comparison of the Experimental Results with Webber's Theory

The experimental results described may be compared with those obtained analytically by Webber by expressing them in the normalized parameters used in the theory. The theoretical results have been computed for a number of different values of $\gamma_0 b$, and in the experimental work the beam diameter was fixed so that the variation in $\gamma_0 b$ corresponds to a variation in the beam accelerating voltage.

The other important parameter is $\omega_q/\omega\alpha$, where ω_q is the reduced plasma angular frequency. This is very similar to the plasma frequency obtained for the first mode in the Ramo and Hahn theory. Webber obtains ω_q from the expression

$$\omega_q = \frac{\omega_p}{\sqrt{\left[1 + \left(\frac{2}{\gamma_0 b}\right)^2\right]}}$$

In Ramo's space-charge-wave theory the effective plasma frequency for the first mode is given by

$$\omega_q' = \frac{\omega_p}{\sqrt{1 + \left(\frac{T_1}{\gamma_0}\right)^2}}$$

In this case the value for Tb is found from an expression involving the beam and drift tube diameters and γ_0. In the special case where the drift tube is infinitely large, $T_1 b = 2{\cdot}4$, and for other values of tube diameter the value is slightly less. From this it is seen that the value of ω_q used by Webber is very similar to that of the first mode of the space-charge-wave theory. For any given beam charge density, ρ_0, this value is fixed and the normalized term is then only affected by variations in the modulation coefficient, α. By evaluating ω_q for the various combinations of beam current and voltage used, the coefficient α may be replaced by $\omega_q/\omega\alpha$.

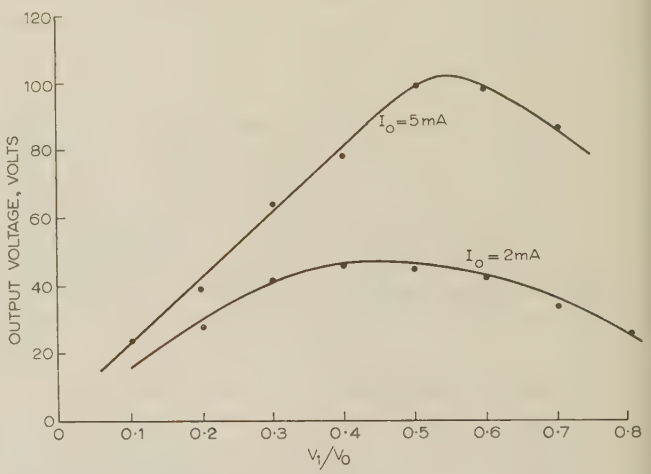


Fig. 3.—Variation of output with V_1.
The drift length is adjusted to be the optimum for small-signal conditions.
$V_0 = 200$ volts.

The evaluation of the r.f. current in the beam from the measured output-cavity voltage is a little more complicated, as a gridless output gap was used. The coupling between any part of the electron beam and the output gap depends on the ratio of the radius of that part of the beam to the gap radius. If a uniform radial distribution of r.f. current in the beam is assumed, an average gap factor may be found. In practice the radial distribution will seldom be uniform, and in analysing the small-signal experiments, space-charge-wave theory was used to find the radial variations and calculate the effective gap factor. The gap factor for each mode was found and the resultant output summed for all modes. As the space-charge-wave theory is

only applicable to small values of $(\omega/\omega_q)\alpha/2$, these gap-factor calculations may be inaccurate for some of the results under consideration. The results are thus plotted as output-gap current rather than r.f. beam current. In comparing these results with those of Webber it must be remembered that the actual r.f. beam current will be larger than the gap current except in very peculiar circumstances. The experimental conditions were such that the first space-charge mode was predominant, and as this has a fairly simple radial distribution the above statement is correct. In the case where the higher-order modes are significant, or at a point where the first mode has a node, it would be possible to have a gap factor greater than unity.

A further difficulty is that the gap current is obtained from the output voltage and the shunt resistance of the cavity. The shunt resistance is affected by the electron-beam loading, and this was allowed for by measuring the cavity Q-factor under the different conditions of beam loading used in practice. It is quite probable that the large depths of modulation used affected the beam loading. In some of the experiments the r.f. voltage appearing across the cavity was greater than the beam direct voltage, so that the small-signal beam-loading correction would obviously be incorrect.

In Fig. 4 one of the experimental curves has been replotted using Webber's normalized parameters. A cursory comparison

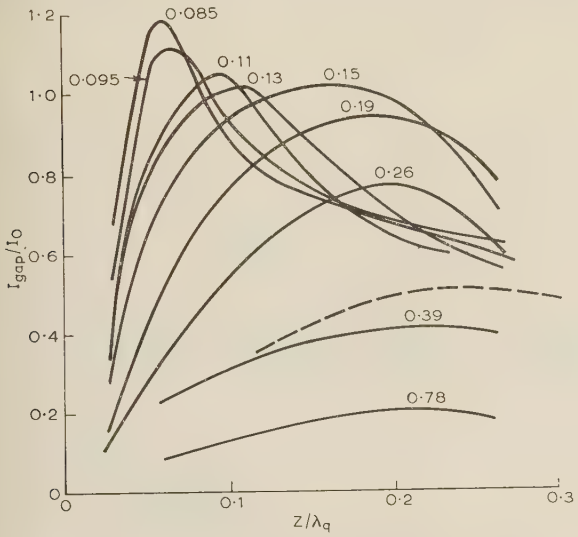


Fig. 4.—Variation of normalized gap current with normalized drift length for various values of $\omega_q/\omega\alpha$.

$V_0 = 200$ volts, $I_0 = 5$ mA, $\gamma_0 b = 2{\cdot}24$

The values of $\omega_q/\omega\alpha$ are shown on the curves. The dotted curve has been calculated from space-charge-wave theory for $\omega_q/\omega\alpha = 0{\cdot}39$.

of this curve with those of Webber shows that both the outputs and optimum drift lengths are similar. One curve on the graph shows the current that would be obtained from multi-mode space-charge-wave theory. This curve allows for the gap factor and so may be compared directly with the experimental curve. Fig. 5 shows the normalized optimum drift lengths plotted against $\omega_q/\omega\alpha$, and Fig. 6 shows the same values obtained from theory. Although there is some divergence in the curves the general variations of drift length with modulation are similar. Fig. 7 shows how the theoretical maximum values of normalized beam current vary with $\omega_q/\omega\alpha$. Fig. 8 shows how the experimental values of normalized gap current vary with the same parameter. In order to estimate the actual beam current from the gap current, the gap factor appropriate to the first mode of the space-charge-wave theory was used. This value was very similar to that obtained by dynamical considerations when the radial distribu-

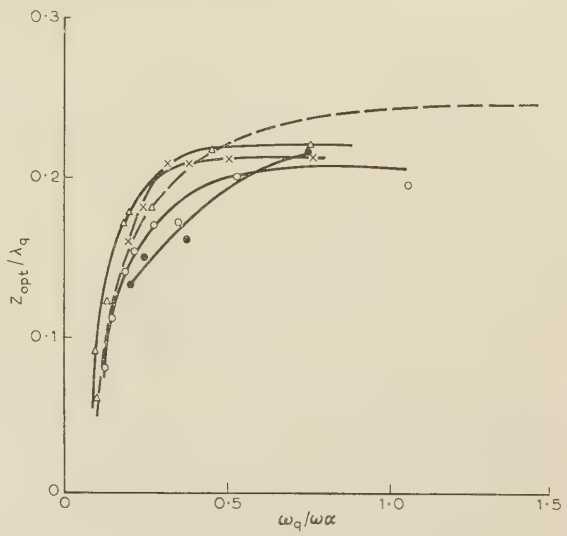


Fig. 5.—Variation of optimum normalized drift length with $\omega_q/\omega\alpha$.

	V_0 volts	I_0 mA	$\gamma_0 b$
○	400	10	1·59
×	200	20	2·24
●	200	10	2·24
Δ	200	5	2·24

The dotted curve is the theoretical curve for $\gamma_0 b = 1{\cdot}6$.

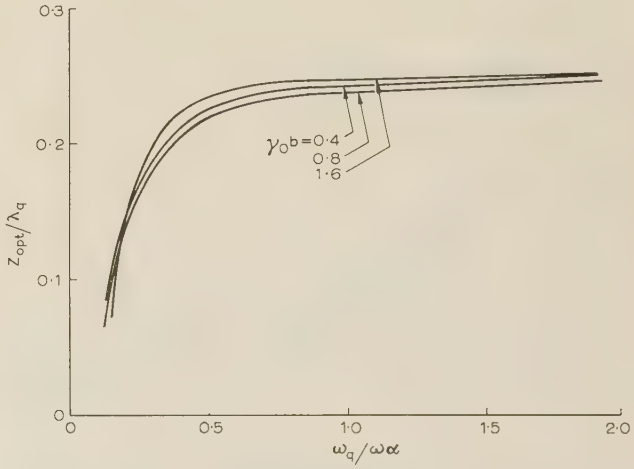


Fig. 6.—Variation of z_{opt}/λ_q with $\omega_q/\omega\alpha$ (Webber's theory).

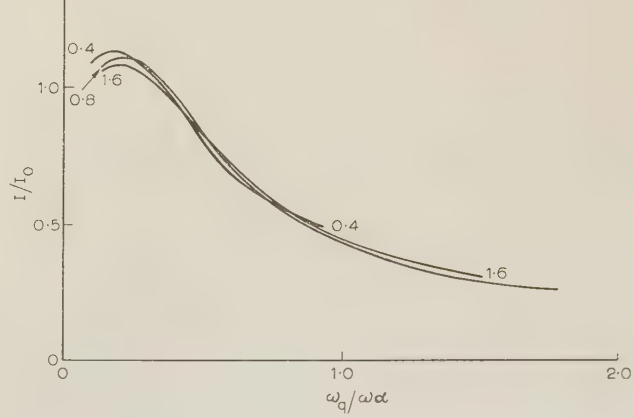


Fig. 7.—Variation of maximum normalized output with $\omega_q/\omega\alpha$ (Webber's theory).

Numerical data used to plot these curves were taken from Figs. 6, 7 and 8 of Reference 4.

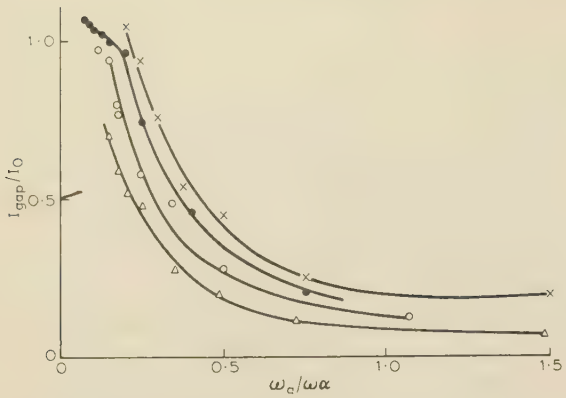


Fig. 8.—Measured values of $\frac{I_{gap}}{I_0}$ (maxi mum) for various values of $\omega_q/\omega\alpha$.

	V_0 volts	I_0 mA
×	200	20
○	200	10
●	200	5
Δ	100	10

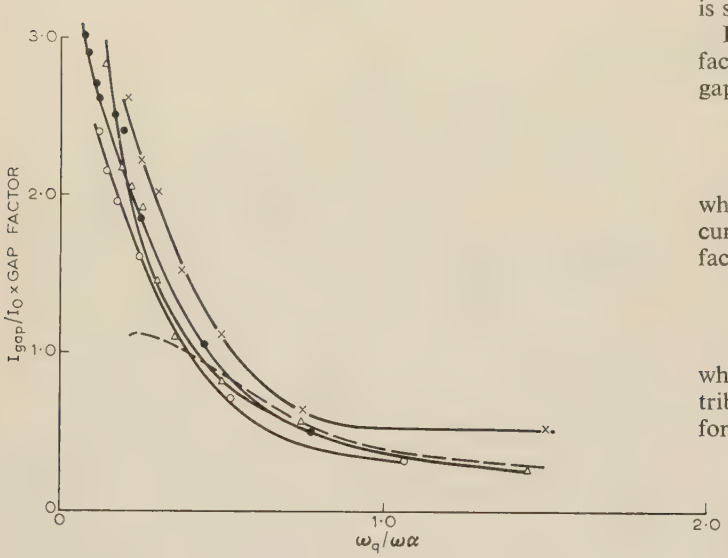


Fig. 9.—Measured values of $\frac{I_{gap}}{I_0 \times \text{gap factor}}$ (maximum) for various values of $\omega_q/\omega\alpha$.

	V_0 volts	I_0 mA
×	200	20
○	200	10
●	200	5
Δ	100	10

tion of r.f. beam current was assumed to be uniform. The estimated value of maximum beam current is shown in Fig. 9. The interesting feature of this curve is that the theoretical limit to the r.f. beam current of $1{\cdot}16I_0$ (according to Webster's kinematic theory) is apparently exceeded by a factor greater than 2 for small values of $\omega_q/\omega\alpha$. The maximum r.f. current that could occur under any condition is $2I_0$. If it is assumed that the theory gives at least an approximately correct value for the maximum current, the error must lie in either the assumed value of cavity shunt resistance or in the estimated gap factor. In the case of the former the indication is that the shunt resistance of the output cavity was higher than the measured value (which was measured in the presence of the beam loading). If the beam-loading effect were removed altogether this would alter the resistance of the output cavity by a factor varying from 1·18 when $V_0 = 400$ to 1·7 at $V_0 = 100$. As the beam loading would only be modified and not eliminated, the effect would be less serious than this. Another source of error lies in the assumed gap factor, If the gap factor were improved to unity, the error would vary from a factor 1·7 to 4·0 for the same two limits of beam voltage.

A simple method of checking this assumption would be to repeat the experiments using a gridded gap, when the gap factor and beam loading would be unaffected by the radial distribution. Unfortunately, tests with gridded gaps showed that secondary-emission loading of the output cavity was very severe and very variable so that reliable results could not be obtained. The variations were probably due to variations in the surface condition, and hence the secondary-emission coefficient, of the grid wires.

(4.3) Gap-Factor Variations

If it is assumed that the alternating component of beam current does not exceed about $1{\cdot}2I_0$, then the apparent improvement of the gap factor can be estimated from the curves of Fig. 9. For values of $\omega_q/\omega\alpha$ near 0·125, the apparent r.f. beam current is slightly more than twice the theoretical value.

In the form of output gap used in the experiments the gap factor at the drift-tube wall was very nearly unity, and so the gap factor at any point at a radius r is given by

$$\beta_r = \frac{I_0(R)}{I_0(R_1)} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where R, R_1 are normalized values. If the r.f. distribution of current is assumed to be uniform, the average or dynamical gap factor is given by

$$\beta = \frac{2I_1(\gamma_0 b)}{I_0(\gamma_0 a)\gamma_0 b} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where I_1, I_0 are Bessel functions. For the case where the distribution is not uniform the radial variations may be allowed for to give a modified form of the above expression.

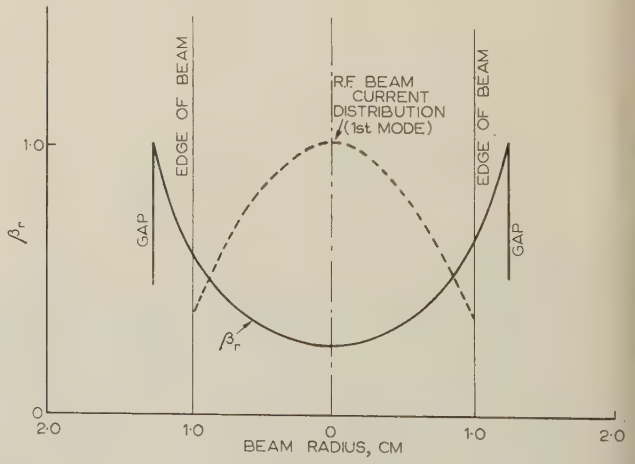


Fig. 10.—Variation of r.f. current and gap factor with radius.

Fig. 10 shows how the gap factor varies with the radius for a beam voltage of 200 volts and it also shows the radial variation of beam current for that voltage. In this particular case it is apparent that the overall gap factor would improve if the current distribution were such that most of the current was carried at the edge of the beam rather than at the centre. In Table 1 the gap

Table 1

OVERALL GAP FACTOR, EVALUATED BY VARIOUS METHODS

b	V_0	5-mode theory	1st mode only	Dynamical	Current at edge	Correction factor
3·17	100	0·26	0·24	0·28	0·51	2
2·24	200	0·40	0·40	0·43	0·67	1·7
1·83	300	0·50	0·50	0·54	0·73	1·5
1·58	400	0·60	0·60	0·60	0·76	1·3

factor has been evaluated by a number of methods, the last of which assumes that all the r.f. current is in a very thin tube having the same radius as the beam itself. Although this is not a practical condition it does serve to give an indication of the improvement in the gap factor that could result from a redistribution of the r.f. beam current.

(4.4) Effect of the Finite Magnetic Field

Webber's theory, like that of Ramo and Hahn, assumes that the axial magnetic field is infinitely large, whilst in practice the field is generally just sufficient to prevent any serious deformation of the beam under normal operating conditions. When the beam is heavily modulated the space-charge density will be large at certain points along the beam, and it is possible that variations will occur both in the radial distribution of current and in the actual shape of the beam. Zitelli,[9] Rigrod and Lewis,[10] and others have considered the case of a klystron with a finite focusing field with a shielded gun, whilst Brewer[11] has extended the work to consider the effect of a finite field linking the electron gun. The results of their work show that in the finite field case the majority of the r.f. current is travelling on the outside of the beam, whereas it was assumed when calculating the gap factor that the distribution was as shown in Fig. 10. This redistribution, due to the finite field, would make the gap factor have a value higher than that used in evaluating the experimental results.

A finite magnetic field would also result in periodic variations in the beam diameter, such that the position at which the maximum diameter occurs coincides with that at which the output maximum occurs. This would again tend to increase the gap factor. The combination of these two effects would partially account for the apparent discrepancies between the theoretical and practical results.

(5) CONCLUSIONS

It is seen from the experimental results that Webber's theory can be used to give a very useful estimate of klystron behaviour under large-signal conditions, where 'large signal' is understood as referring to the value of $\omega_q/\omega\alpha$ rather than to α. When making a numerical comparison it must be remembered that theory is subject to limitations imposed by:

(*a*) The assumption of sinusoidal velocity modulation at the input gap. This will not be true for voltages large compared with the beam accelerating voltage, as the velocity is proportional to $\sqrt{(V_0 + V_1)}$ and it is the voltage that is sinusoidal.
(*b*) The neglect of radial variation in r.f. current.
(*c*) The assumption of an infinite focusing field.

As Webber points out in his original work, the inclusion of only one of the above points, that of radial variations, would make the computation virtually impossible on present-day computers and it would be even more difficult to include the other factors. Whilst a more complete analysis would be interesting, it is doubtful whether the extremely laborious computations would be of very great practical use, as the geometrical arrangement in practical klystron would probably depart from that of the ideal model and such factors as the radial distribution of the d.c. beam would probably cause a greater numerical discrepancy than is involved by using a simpler theory.

The experimental results may also differ from the calculations for the following practical reasons:

(*a*) The output gap factor could not be accurately evaluated, as the radial variations of r.f. current were not known.
(*b*) In many cases the output-gap voltage was rather large compared with the beam voltage.
(*c*) Numerical results obtained from Webber's theory are only available for a limited range of values of $\gamma_0 b$.

For most practical design purposes the optimum drift length is of more interest than the actual output current, and it is seen that the reduction of the optimum drift length at high signal levels is predicted with reasonable accuracy by the theory.

A further important conclusion from both the experiments and the theory is that the parameter that determines whether or not the signal is large is $(\omega/\omega_q)\alpha/2$ and not merely α.

Experiments with variable-drift-length klystrons have also been performed by Chodorow[6,7] and his colleagues and by Mihran.[8] The latest results reported by Chodorow[7] are of interest as they demonstrate the radial distribution of r.f. current under small-signal conditions. It is seen from the results that quite a large proportion of the total r.f. current is carried by the edge of the beam and the distribution is affected by the magnetic focusing field strength. As the results are for a valve with a shielded electron gun, they cannot be compared directly with those given here, but in view of Brewer's remarks on the presence of a finite field at the cathode in most practical valves, they are of interest.

Mihran[8] has made an investigation using a klystron with an immersed electron gun. His results include large-signal operation, and it is again seen that the optimum drift length is considerably reduced when the input signal is large. Mihran's results show the output power as a function of the drift length for various input power levels, and the shape of the curves is very similar to those shown in Fig. 1. There are not sufficient data in the paper to convert these input and output power levels into terms of the input modulation depth and output r.f. current, but the change-over from the small-signal conditions, where space charge predominates, to the large-signal condition is again clearly marked.

(6) ACKNOWLEDGMENTS

This work was performed in the Electrical Engineering Laboratories at University College, London, under the supervision of Professor H. E. M. Barlow, to whom the author is grateful for much help and encouragement. The experimental work was originally suggested by Professor A. L. Cullen, to whom the author is grateful for constant help and advice, both during the course of the experimental work and in the preparation of the paper.

Thanks are due to the Admiralty for permission to publish the paper.

(7) REFERENCES

(1) WEBSTER, D. L.: 'Cathode Ray Bunching', *Journal of Applied Physics*, 1939, **10**, p. 501.
(2) RAMO, S.: 'Space Charge and Field Waves in an Electron Beam', *Physical Review*, 1939, **56**, p. 276.
(3) HAHN, W. C.: 'Small Signal Theory of Velocity Modulated Electron Beams', *General Electric Review*, 1939, **42**, p. 258.
(4) WEBBER, S. E.: 'Ballistic Analysis of a Two-Cavity Finite Beam Klystron', *Transactions of the Institute of Radio Engineers*, 1958, **ED–5**, p. 98.

(5) CULLEN, A. L., and STEPHENSON, I. M.: 'An Experimental Investigation of Velocity-Modulated Electron Beams', *Proceedings I.E.E.*, Paper No. 2472 R, May, 1958 (**105** B, Suppl. No. 12, p. 843).
(6) CHODOROW, M., GINTZEN, E. L., and NALOS, E. J.: 'Debunching of Electron Beams Constrained by Strong Magnetic Fields', *Proceedings of the Institute of Radio Engineers*, 1953, **41**, p. 999.
(7) CHODOROW, M., SHAW, H. J., and WINSLOW, D. K.: 'Current Distribution in Modulated Magnetically Focused Electron Beams', *Journal of Applied Physics*, 1958, **29**, p. 1525.
(8) MIHRAN, T. G.: 'The Effect of Space Charge on Bunching in a Two-Cavity Klystron', General Electric Research Laboratory Report, 1958, No. 58–RL–1916.
(9) ZITELLI, L. T.: 'Space Charge Effects in Gridless Klystrons', Microwave Laboratory Report No. 149, Stanford University, 1951.
(10) RIGROD, W. W., and LEWIS, J. A.: 'Wave Propagation along a Magnetically-Focused Cylindrical Electron Beam', *Bell System Technical Journal*, 1954, **33**, p. 399.
(11) BREWER, G. R.: 'Some Effects of Magnetic Field Strength on Space-Charge-Wave Propagation', *Proceedings of the Institute of Radio Engineers*, 1956, **44**, p. 896.

THE USE OF A WIRE-WOUND HELIX TO FORM A CIRCULAR H_{01} WAVEMETER CAVITY

By Prof. H. E. M. BARLOW, Ph.D., B.Sc.(Eng.), Member, H. G. EFFEMEY, Associate Member, and P. H. HARGRAVE, B.Sc.

Cavity wavemeters for microwave measurements frequently employ the circular H_{01} mode, because this enables a non-contact piston to be used without a choke joint. Unfortunately the circular H_{01} mode is not the dominant one in a smooth-bore metal tube, and, in consequence, difficulties often arise from the excitation of other modes. To suppress unwanted modes special arrangements are sometimes made in the design of the piston, using, for example, a concave end, but, even so, spurious field distributions inside the cavity are not uncommon, and they can lead to serious inaccuracies in measurements with the instrument.

Experiments have shown that, if the cylindrical cavity [Fig. 1(*a*)] is formed from a wire-wound helix of pitch small compared with the wavelength, the Q-factor is only very slightly reduced while all modes other than the H_{0n} family are almost completely suppressed. This is basically the same technique as that used by Miller* for the low-loss waveguide application, and the construction of the helix for the wavemeter was developed in a similar way. Thus, for an H_{01} type of cavity wavemeter applied to the 8 mm band and having a cylindrical channel about 0·68 in inside diameter, No. 34 s.w.g. enamelled copper wire was wound with adjacent turns in contact on a stainless-steel mandrel. The assembly was then placed along the axis of a rather larger tube, leaving an annular space between its inside surface and the helix, of about $\frac{1}{4}$ in radius. After filling this space with a suitable plastic (e.g. epoxy resin) and allowing it to set, the outer tube was removed and the stainless-steel mandrel was withdrawn. Although there is a detectable field outside the helix when the cavity is excited, it decays very rapidly with increasing radius and is not a significant factor in the behaviour of the wavemeter, when ordinary thicknesses of enamel covering on the wire are employed and the spacing between adjacent turns of the helix is correspondingly small.

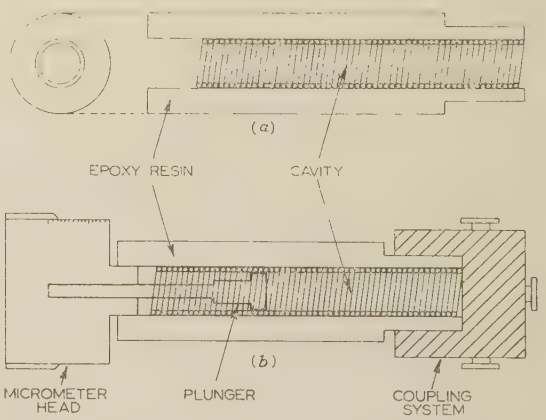


Fig. 1.—Helical wire structure for cavity wavemeter.
(*a*) Helical wire cavity.
(*b*) Schematic of helix wavemeter.

A conventional wavemeter, adapted [Fig. 1(*b*)] to use a helical wire cavity of this kind, has been found to give an exceedingly high degree of mode purity and to have a greatly improved performance.

* MILLER, S. E.: 'Electromagnetic Wave Transmission', U.S. Patent No. 2 848 696, August, 1958.

Prof. Barlow is Pender Professor of Electrical Engineering, University College, London.
Mr. Effemey and Mr. Hargrave are in the Electrical Engineering Department, University College, London.

621.317.794

The Institution of Electrical Engineers
Paper No. 3146 E
Jan. 1960
©

A COAXIAL FILM BOLOMETER FOR THE MEASUREMENT OF POWER IN THE U.H.F. BAND

By I. A. HARRIS, Associate Member.

(*The paper was first received* 23*rd April, and in revised form* 30*th July*, 1959.)

SUMMARY

The requirements to be satisfied by a transfer element for the comparison of u.h.f. with d.c. power are stated under the headings of transfer properties and matching and residual losses, and the need for a new type of coaxial bolometer is discussed in the light of them. A coaxial film bolometer which satisfies these requirements for all frequencies between 200 Mc/s and 4 Gc/s is described, and the factors affecting its accuracy are discussed. The range of power that can be handled directly by the instrument is 2–200 mW, the v.s.w.r. relative to 50 ohms is of the order of 0·98 and the uncertainty in measurement is within 1% for powers between 20 and 200 mW.

The design is not confined to a particular size. A smaller bolometer element, which handles powers down to 0·1 mW with an upper frequency limit of about 20 Gc/s, and a larger bolometer element, which handles powers up to 2 watts with an upper frequency limit of 3 Gc/s, are feasible.

LIST OF SYMBOLS

C = Capacitance, F.
I = Direct current to bridge, amp.
P = Power, watts.
R = Feed resistance, ohms.
ΔR = Incremental feed resistance, ohms.
R_a = Bridge arm resistance, ohms.
R_L = R.F. load resistance, ohms.
R_g = Galvanometer resistance, ohms.
s_G = Source voltage standing-wave ratio.
s_L = Load voltage standing-wave ratio.
α = Temperature coefficient of resistance per deg C.
β = Bolometer sensitivity, ohms/watt.
ϵ = Fractional error.
θ = Temperature rise, deg C.
ω = Angular frequency, rad/s.

(1) INTRODUCTION

Under suitable conditions, d.c. electric power dissipated in a resistor can be measured with a high degree of precision, with the result that several methods of measuring r.f. power in terms of d.c. power by the employment of some form of resistive 'transfer element' have been devised. For example, a resistor that has substantially the same resistance to r.f. as to direct current, together with means for detecting equality in the rise of its temperature when heated either by r.f. or d.c. power, may be used. Arrangements in which both r.f. and d.c. power may be applied to the resistor simultaneously have advantages, particularly for the precise measurement of a few milliwatts. To detect equality of total power irrespective of r.f./d.c. composition, a resistor with a high temperature coefficient of resistance is employed as a sensitive resistance thermometer as well as a dissipative element. In the d.c. circuit, this forms one arm of a Wheatstone bridge, the three other arms of which are resistors which do not change appreciably in resistance when the direct current is varied. The bridge, which is fed with measured d.c. power, is balanced when the total power in the transfer element is at a given level. R.F. powers of less than this level are determined by their substitution for part of the initial d.c. power in the element.

A transfer element used in the manner described is usually given the name *bolometer*, and the particular method of use is known as the substitution method. This is considered to be the most suitable method for a precision instrument of the type to be described. Because there is no direct experimental way to determine the effectiveness of a transfer element for precision work, other than by the consistency of results obtained by equally-sound alternative methods, it is advisable to study the action of each element in detail. This can reveal those factors which may result in error. Accordingly, before describing the coaxial film bolometer, the requirements of a satisfactory transfer element and how they are met by some other types of bolometer are briefly discussed.

(2) CONDITIONS TO BE SATISFIED BY A TRANSFER ELEMENT

The degree of accuracy attained by any form of bolometer depends, first, on how closely the transfer element satisfies the condition that the d.c. resistance is the same function of r.f. power as it is of d.c. power; secondly, on how closely the bolometer mount matches the nominal resistive impedance for which it is designed; and thirdly, on the proportion of the power accepted at the inlet to the mount that is dissipated in the transfer element.

(2.1) Transfer Properties

The presence of a standing wave of r.f. current along the length of an element results in different temperature distributions for r.f. and d.c. heating. Thus with an element in which the heat equilibrium is governed by radiation and end conduction (such as a tungsten filament *in vacuo*), or by radiation, convection and end conduction (such as a wire of moderate thickness or a metal film in air), it has been shown[1, 2] that, for a given change in d.c. resistance, the change in r.f. power differs appreciably from the change in d.c. power. The difference is a matter of degree, but nevertheless, even when the length of wire is less than 0·1 wavelength, it is not negligible. On the other hand, a short length of extremely thin (0·000 04 in) platinum wire (Wollaston wire) in air has the property that the power density and change in resistivity are nearly linearly related at all parts of the wire. As a result, the total change in d.c. resistance depends very nearly on the total change in power, irrespective of the distribution.[3] Because of the approximate nature of the linearity there is a limit to the ratio of length to wavelength which can be permitted for a given transfer error.

Another approach to the problem is to make the relevant properties of the element the same for r.f. as for d.c. power. Thus the following should be the same for both types of current:

(*a*) Distribution and magnitude of resistance.
(*b*) Distribution of current.

* Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

† Mr. Harris is at the Laboratories of the Electrical Inspection Directorate, Ministry of Aviation.

Means for the attainment of these two conditions will be discussed in later Sections.

(2.2) Matching

The most useful definition of the output power of any radio-frequency (v.h.f./u.h.f.) generator is the power absorbed by a non-reactive load of given resistance, and it is essential that the inlet to the bolometer mount should have an impedance that matches this nominal resistance over as wide a frequency band as possible. The nominal impedance, as the load resistance is termed, is usually equal to the nominal characteristic impedance of the transmission line from the generator. In practice, both the source impedance of the generator and the load impedance of the power meter differ from the nominal impedance. Let this difference be expressed in each case by the voltage standing-wave ratio relative to the nominal impedance. If s_G is the source v.s.w.r. and s_L is the load v.s.w.r., it can be shown that the upper and lower limits of the ratio, P/P_0, of the power actually absorbed by the load to the power that would be absorbed by the nominal impedance are as follows:

$$\left.\begin{aligned}(P/P_0)_{max} &= s_L(1+s_G)^2/(s_L+s_G)^2\\ (P/P_0)_{min} &= s_L(1+s_G)^2/(1+s_Ls_G)^2\end{aligned}\right\} \quad . \quad . \quad (1)$$

When $s_G = 1$, the two limits coincide to form a single correction factor:

$$P/P_0 = 4s_L/(1+s_L)^2 \quad . \quad . \quad . \quad . \quad . \quad (2)$$

When $s_L = 1$, the ratio is unity irrespective of s_G.

The source mismatch can often give $s_G = 0{\cdot}5$, and if the v.s.w.r. of the power meter is $s_L = 0{\cdot}9$, the resulting uncertainty is $\pm 3{\cdot}7\%$ centred on the value given by eqn. (2). With $s_G = 0{\cdot}5$ and $s_L = 0{\cdot}98$, the uncertainty is $\pm 0{\cdot}7\%$, which is just acceptable for precision work. It is apparent that precision in matching a bolometer to the nominal impedance is important, unless the bolometer is to be restricted to working with a closely matched source having $s_G > 0{\cdot}9$.

(2.3) Residual Losses

Residual losses in a bolometer mount are usually of two kinds—the resistance of the inlet conductors with any fixed or variable matching devices, and dielectric losses associated with any insulating support or envelope surrounding the bolometer element. Under certain conditions the losses can be measured directly.[4, 7]

(3) NEED FOR A NEW TYPE OF BOLOMETER

(3.1) Existing Types of Bolometer Mount with a Coaxial Inlet

One form of a very broadband (10 : 1) mount is the 'disc' mount[5] in which a mica disc supports two equal bolometer elements, either Wollaston wire or resistive film, so that, for r.f. current, they bridge the ends of the inner and outer conductors of a coaxial transmission line on opposite halves of a diameter. Built-in blocking capacitors allow the formation of a d.c. circuit through the two elements in series. The 'series d.c.–parallel r.f.' arrangement eliminates the need for a d.c. return path across the r.f. inlet. A somewhat different mount[8] employing this arrangement has been used with a pair of bead thermistors* as the elements. Unlike other bolometer elements, bead thermistors are not simple resistive elements. Experiments[7] on the measurement of their efficiency as transfer elements at frequencies of 3 Gc/s and higher have shown that it is doubtful whether the bead resistance retains its d.c. value at frequencies higher than about 2 Gc/s. Other experiments have shown that at lower frequencies the r.f. and d.c. resistances of the bead differ by less than 1%, and at 200 Mc/s the difference is within 0·2%.

The v.s.w.r. obtained with most broadband coaxial mounts varies over the frequency band from about 0·75 to more than 0·9. The use of a tunable mount, or a coaxial mount with a single element and the inevitable shunt resonant stub across the r.f. inlet to provide a d.c. return path, can result in a better v.s.w.r. over a narrow band; but the sliding parts of the tunable mount or the stub of the single-element mount are liable to introduce losses[2] which necessitate the measurement of the mount efficiency.

(3.2) Waveguide Bolometer Mounts

Most waveguide mounts employ either a bead thermistor or an encapsulated Wollaston wire as the element. Experiments[10] with a newer type of waveguide bolometer which employs a narrow strip of platinum film deposited on mica, mounted parallel to the electric field in the centre of the waveguide, have shown that the transfer error of the film at 9 Gc/s is within 2%, in contrast to the −10% of the bead thermistor (arising partly from the properties of the bead and partly from mount losses) and of the encapsulated Wollaston wire[7] (arising from mount losses). Under optimum conditions a v.s.w.r. of better than 0·9 has been obtained. Since the film is certain to lose appreciable heat by radiation as well as convection, the experimental results suggest that conditions (*a*) and (*b*) of Section 2.1 are satisfied, at least to a close approximation.

(3.3) Need for an Improved Bolometer with a Coaxial Inlet

Among the conditions to be satisfied by a bolometer, that of matching over a wide band of frequencies has not been met closely by existing designs. Admittedly, the error in power measurement resulting from the use of a poorly-matched bolometer head can sometimes be reduced to a negligible value by employing a source with a v.s.w.r. of better than 0·9, but this is not always convenient or possible. There is therefore a need for a new form of coaxial bolometer which satisfies conditions (*a*) and (*b*) of Section 2.1, as well as the conditions of matching and low residual loss, over the whole of the u.h.f. band.

A greatly modified version of the original 'series d.c.–parallel r.f.' thermistor mount[8] has been developed by the author, which has a v.s.w.r. of better than 0·97 over the frequency band 100 Mc/s–1·25 Gc/s with a transfer error within 1% and a range of 20 μW–2 mW. However, the frequency range is insufficient for all u.h.f. measurements.

As a result of the success of Lane's waveguide film bolometer,[10] and of a particular form of coaxial resistor mount,[11] the possibility of adapting the resistor mount to the design of a coaxial film bolometer was studied. The outcome of the study was a coaxial film bolometer that satisfied inherently the conditions for a precise transfer element.

(4) COAXIAL FILM BOLOMETER

(4.1) General Description

An axial section of the bolometer head is shown in Fig. 1. The element for dissipating the r.f. power comprises a coaxial line having an inner conductor in the form of a cylindrical resistive film with an outer sheath of copper, the axial section of which over the length of the resistor has the form of a tractrix. The inner and outer lead-in conductors are in the form of a uniform conical coaxial line, the outer cone being an extension of the tractorial form. An arrangement of aligned, thin slits in the film resistor and in the outer sheath separates these into two parts each (Figs. 1, 2 and 3). The two half-flanges forming

* A bolometer using a thermistor as the element is not usually described in this country as a bolometer, but as a thermistor or a thermistor bridge. This usage probably stems from Definition No. 1562 of B.S. 204: 1943, which implies that a bolometer element is restricted to 'a fine wire, strip or filament'.

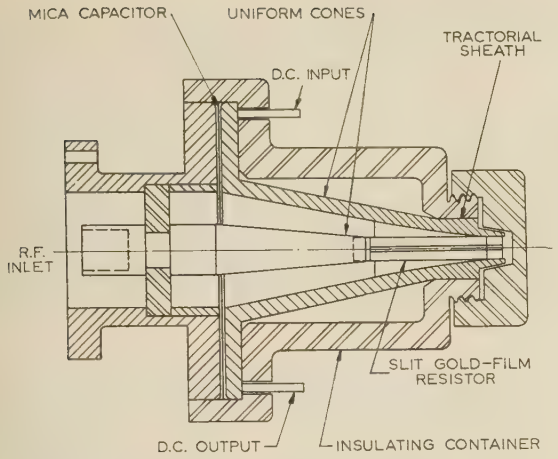


Fig. 1.—Axial section of the film-bolometer head.

extensions of the two parts of the tractorial sheath and lead-in cone are used in conjunction with a mica ring and a flange on the outer inlet conductor to form a d.c. blocking capacitor. The slits and the blocking capacitor permit direct current to be fed through the two halves of the film resistor in series via the two halves of the outer sheath and the half-flanges. The slits are thin (up to 0·01 in) and are parallel to the paths of the r.f. currents in the resistor and the sheath, so that they do not interfere appreciably with the r.f. properties of the resistor mount.

The inner inlet cone is hollow at the smaller end to take the silvered end of the film resistor. Two thin sawcuts are made to facilitate clamping the resistor. The larger end of this inner cone is secured by the p.t.f.e. support, which is clamped in undercut sections of the inner and outer inlet lines in such a way that the discontinuity in characteristic impedance arising from the presence of the support is compensated. The r.f. power enters through the coaxial inlet as a transverse guided electromagnetic wave and is propagated without reflection through the insulating support and the conical line to the film resistor, where it is uniformly and completely dissipated.

(4.2) Method of Construction

The 50-ohm film resistor (Fig. 2) comprises a thin hard-glass cylinder of low dielectric loss, which has been ground to ensure

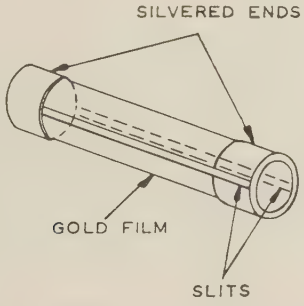


Fig. 2.—Detail of the slits in the film resistor.

that the outer diameter does not depart from the nominal 0·365 in by more than 0·001 in. A gold film is deposited on the outer surface with a surface resistivity of about 32 ohms per square. To obtain uniformity, the cylinder must be rotated while the deposition is taking place. Bands of silver are deposited around each end of the cylinder to provide adequate contacts and to define the length of the gold film to within 0·01 in. To form the slits, the resistor is clamped by the ends in a jig which allows connections to be made, two opposite connections at one end and one at the other. Thin slits are scored in each side of the resistor so that the resistance of each half is 99·3 ohms $\pm$ 0·2 ohm at room temperature.

The outer sheath is made by electro-deposition of copper on a stainless-steel former. A brass flange is soldered to the larger end and the assembly is sawn across a diameter to form two

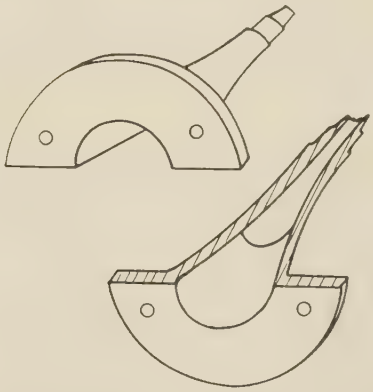


Fig. 3.—Split outer conductor and flange.

pieces, as shown in Fig. 3. These two pieces are separated by shaped strips of insulating material to replace the metal sawn away, and the whole is securely clamped together by a cylinder of insulating material. The nut used to clamp the end of the resistor remote from the inlet is also made of insulating material.

(4.3) Resistor Mount and the Inlet Conductors

It has been shown[11] that, provided that there is substantially complete penetration of r.f. current through the film, the tractorial form of outer conductor can be designed so that the impedance at the coaxial inlet to the resistor is equal to its d.c. resistance with no reactance, to a high degree of precision. This applies at all frequencies up to that at which supplementary modes of propagation occur in the inlet line. Furthermore, the distribution of current in the resistor is uniform with either d.c. or r.f. heating. As a result of end cooling, however, a temperature gradient exists at each end; but, in the present bolometer element, the resultant variation of resistivity is 0·7% at the most and has no appreciable effect on the r.f. current distribution. It follows that conditions (*a*) and (*b*) of Section 2.1 are satisfied together.

The uniformity of the r.f. current distribution is explained as follows. With a cylindrical outer conductor having a tractorial axial section, the amplitudes of the electric and magnetic field intensities do not vary along the mount.[11] The energy per unit volume of the field at any place between the resistor and its sheath is therefore constant, and this energy moves with the velocity of wave propagation along and slightly into the resistor, as indicated in Fig. 4. By reason of the shape of the sheath, the volume of a given short length of the field (AB in Fig. 4) is reduced at a uniform rate as it moves along the system. The energy of the 'lost' volume of field is converted into heat in the resistor, which is therefore uniformly heated by the passage of the wave. When the short length of field considered arrives at the end of the resistor remote from the inlet, its energy has all been converted into heat and none is left for reflection. It should be noted that the voltage between the sheath and the resistor is zero at this end; but the current is constant in ampli-

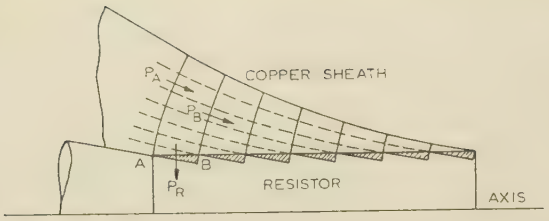


Fig. 4.—Representation of the uniform heating by an electromagnetic wave of a resistor mounted in a tractorial outer conductor.

The electric and magnetic flux densities remain constant in amplitude, while the wavefront area is reduced as the wave propagates from left to right. The area of each shaded triangle is proportional to the equivalent amount of field converted into heat.

Note that $P_R = P_A - P_B$.

tude along the system and must have a zero-impedance path at the end. The smooth process of uniform dissipation ensures no reflection of the wave at any point along the resistor, and in the absence of reflection there is no reactive component of impedance at the inlet.

To obtain a v.s.w.r. of 0·98 over the u.h.f. band, it is essential that the profile of the sheath be realized to within 0·001 in. Stepped or uniform conical approximations to the tractorial profile might possibly give an acceptable v.s.w.r., but the uniformity of the r.f. current distribution would be adversely affected to an uncertain degree. It is equally important that the coaxial inlet and the support for the inner conductor maintain a v.s.w.r. of better then 0·98. A connector system developed for precision v.h.f. and u.h.f. measuring instruments[12] is adopted, in which the inner and outer conducting surfaces are of 0·325 7 and 0·750 0 in diameter, respectively.

(4.4) Design of the Mount and the Bolometer Element

The design of the tractorial profile and the lead-in cones has been described elsewhere.[11] Allowance must be made for the finite width of the slits in the resistor, as well as for the field that penetrates the film. The lowest frequency at which the bolometer can be used is governed by the capacitance of the mica blocking capacitor. Thus, if the resistor mount has the substantially perfect resistance R_L, a series capacitance C will result in a v.s.w.r. given approximately by

$$s = 1 - 1/\omega CR_L \quad (\omega CR_L \gg 1) \qquad (3)$$

The measured value of C in the bolometer constructed is 800 pF, which gives $s = 0{\cdot}98$ at 200 Mc/s. By improving the details of the mica capacitor (e.g. by silvering the faces except for appropriate margins and using thinner mica) the capacitance can be increased to 1 600 pF, which will lower the minimum frequency to 100 Mc/s. It is important to place the d.c. blocking capacitor at the entrance to the mount and not at the remote end, because in the latter position it would interfere with the electric-field distribution in the tractorial part of the mount.

The bolometer sensitivity, β, is the rate of change of d.c. resistance with respect to the power dissipated. A simple calculation shows that β varies directly as the film surface resistivity and inversely as the square of the diameter of the resistor cylinder. The minimum length/diameter ratio of the resistor, consistent with satisfactory design of the mount, is 5 : 1 for a resistance of 50 ohms. This limits the maximum film surface resistivity to about 32 ohms per square. With the length/diameter ratio fixed, the power rating for a given temperature rise varies as the square of the resistor diameter. Therefore, with optimum design, the sensitivity varies inversely as the power rating of the bolometer element. However, the resistance change for a given fraction of the total power remains constant.

(4.5) D.C. Circuit and Mode of Use

The r.f. resistance of 50 ohms is formed by the two halves of the film resistor in parallel, while the direct current passes through the two halves in series giving a d.c. resistance of 200 ohms, notwithstanding the equality of the r.f. and d.c. distributions of current density irrespective of direction. This 200-ohm resistance forms one arm of a Wheatstone bridge.

The circuit used is shown in Fig. 5. The Wheatstone bridge

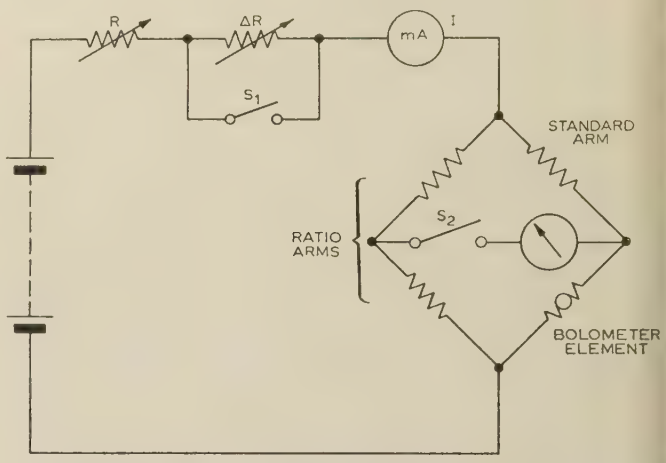


Fig. 5.—Bridge circuit for use with the bolometer head.

is fed by a d.c. source of very low resistance, such as a 27-volt battery of large-capacity secondary cells, through a precision decade resistance box of suitable rating. A second decade resistance box (ΔR in Fig. 5), which can be shunted by a switch of low residual resistance, is also placed in the d.c. feed circuit. The current to the bridge is measured by a precision-grade instrument (0–100 mA). For measurements of r.f. power between 20 and 200 mW, the minimum division of resistance of 0·01 ohm is sufficient, and the galvanometer sensitivity should be 100 mm per μA. For r.f. power levels down to 2 mW, a resistance division of 0·005 ohm and a galvanometer sensitivity of at least 1 000 mm per μA are necessary. In use, the bridge is balanced with direct current alone with the switch S_1 closed. The value of R (the total feed resistance including the box, source and milliammeter) and the current I are noted. The r.f. power is then applied, opening the switch S_1 at the same time. At the first reading, the approximate value of ΔR required to rebalance the bridge is noted. The r.f. power is removed, S_1 is closed and the bridge is rebalanced with direct current before a second reading is made. A final check with the r.f. power switched on and off, with simultaneous switching of S_1, should produce no deflection on the galvanometer. The change in d.c. power is given by

$$\Delta P = \frac{I^2 R_a \Delta R}{2(R_a + R + \Delta R)}\left[1 - \frac{\Delta R}{2(R_a + R + \Delta R)}\right] \qquad (4)$$

where R_a is the resistance of each arm of the bridge (200 ohms).

The thermal time-constant of the resistive film is about 12 sec, but application of the procedure outlined minimizes any change in its temperature, and if a galvanometer of six times the normal sensitivity is used, the required degree of balance can be made in about 4 sec.

(5) SOURCES OF ERROR

(5.1) Errors which are Independent of Power Level

(5.1.1) Mismatch and Uniformity of the Film.

The effect of mismatch has already been discussed. With a v.s.w.r. of not worse than 0·98 the uncertainty in power measurement with a source v.s.w.r. of not worse than 0·5 is ±0·7%.

A small departure from uniformity in the resistive film will not appreciably affect the similarity between the d.c. and r.f. power distributions, provided that the r.f. impedance does not show a marked departure from the nominal resistive value. This observation is true only when the sheath surrounding the resistor is of properly designed tractorial form.

(5.1.2) Residual Losses.

The resistance of the inlet conductors and the sheath surrounding the resistor is about 0·047 ohm at 1 Gc/s, with a frequency dependence proportional to $f^{1/2}$. Compared with the 50-ohm resistance of the resistor, this amounts to a loss in power of 1 part in 10^3 at 1 Gc/s. The losses in the joints can be reduced to a very small value by suitable design and surface finish. The p.t.f.e. support is equivalent in loss to a shunt resistance greater than 1 megohm at 1 Gc/s, with a frequency dependence f^{-1}, and is negligible. The fringe field penetrating the hard-glass cylinder that supports the resistor is a small fraction of the total field, and the loss that this couples into the coaxial system is negligible.

(5.1.3) Variation in the Standard Arm of the Bridge.

The sensitivity of the gold-film bolometer element is 6 ohms/watt, compared with 5 ohms/mW for a Wollaston wire or 10 ohms/mW for a bead thermistor. The low sensitivity of the gold film restricts the permissible change in the standard arm of the bridge to $1{\cdot}2 \times 10^{-3}$ ohm for the change in d.c. power of 200 mW, if the resulting error in measurement is to be 0·1%. This occurs because the resistance of the bolometer element is always balanced against that of the standard arm. The percentage error is independent of the power measured. The requirement can be achieved by the use of manganin wire of size adequate for the standard arm of the bridge.

(5.2) Differences between the two Halves of the Film Resistor

If, as a result of inaccurate division of the uniform film by the slits, one half has resistance R_1 and the other half R_2 under d.c. conditions, it can be shown that the fractional error in the determination of r.f. power, on this account, is given by

$$\epsilon = \frac{-(\beta_1 - \beta_2)(R_1 - R_2)}{R_1\beta_2 + R_2\beta_1 - \beta_1\beta_2(P - 2\Delta P)} \quad . \quad . \quad . \quad (5)$$

In this expression, $\beta = dR/dP$ is the bolometer sensitivity, P is the power in the whole element with d.c. power alone applied and ΔP is the reduction in d.c. power required to rebalance the bridge when r.f. power is also applied. The suffixes 1 and 2 refer to the two halves of the resistor. If, for example, β_1 and β_2 differ by 1% and R_1 and R_2 differ by 1%, then, with the mean values $\beta = 6$ ohms/watt, $R = 100$ ohms, and with $P = 0{\cdot}25$ watt, $\Delta P = 0{\cdot}01$ watt then $|\epsilon| = 5 \times 10^{-5}$, which is negligible.

Metal films have a positive temperature coefficient of resistance (and therefore β), and it follows that the parallel combination of a pair of film resistors is a stable configuration while the series combination is less stable when dissipating power. Since the series combination forms the d.c. path, the values at which R_1 and R_2 settle under the d.c. dissipation can be measured, and the worst condition can thus be known with certainty. The stability of the parallel combination results in a decrease in $|\epsilon|$ as the r.f. approaches the maximum.

(5.3) Errors that Determine the Minimum Power which can be Measured

The comparison of r.f. with d.c. power relies on constant temperature of the bolometer element. Therefore a change in ambient temperature during the brief period in switching from the d.c. condition to the d.c. plus r.f. condition and observing the two conditions will result in error. A falling ambient temperature results in a low result for the transfer d.c. to d.c. plus r.f. condition and a high result for the opposite transfer, while the reverse applies with a rising ambient temperature. If the rate of change of temperature varies little during a measurement and if the intervals between readings are similar, the mean of a pair of reversed transfer readings will have a greatly reduced error. The change $\delta\theta$ in temperature corresponding to a change δP in power in the bolometer element is given by

$$\delta\theta = \beta\delta P/\alpha R_a \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

where the temperature coefficient of resistance, α, is $6{\cdot}4 \times 10^{-4}$ per deg C for the gold film (unlike $3{\cdot}4 \times 10^{-3}$ per deg C for bulk gold), the bolometer sensitivity, β, is 6 ohms/watt and R_a is 200 ohms. The lowest power measured with the present bolometer is 2 mW, and 1% of this corresponds to $\delta\theta = 0{\cdot}001°$ C. To attain the equivalent of this accuracy, even under controlled laboratory conditions, a pair of reversed measurements is essential.

The thermo-electric e.m.f. developed in the bolometer head under normal operating conditions is 1 μV, which is negligible. Assembly of the three fixed arms of the bridge and their junctions in one compact unit, immersed in oil, improves uniformity of temperature and reduces the change in temperature with change in d.c. power.

The galvanometer sensitivity required to ensure a given uncertainty in the lowest power measured can be deduced from the following approximate formula:

$$I_g = \frac{I\beta\delta P}{4(R_a + R_g)} \quad . \quad . \quad . \quad . \quad . \quad (7)$$

The galvanometer sensitivities quoted in Section 4.5 were derived from this formula. The accuracy and stability required in the feed-resistor decades can be ascertained from eqn. (4). Thus with $R = 150$ ohms, $I = 70$ mA and $R_a = 200$ ohms, a resistance change of 0·005 gives a change of 0·35% in the measurement of 2 mW r.f. power.

(5.4) Summary of Errors and Uncertainties

For all r.f. powers up to 200 mW, the following factors affect the uncertainty in measurement:

(i) Mismatch, when the source mismatch is not small (0·7%).
(ii) Constancy of the standard bridge arm resistance (0·1%).
(iii) Residual losses (0·1%).

For r.f. powers between 20 and 2 mW, the following additional uncertainties are appreciable:

(iv) Variation in temperature during a measurement (0·1–1%).
(v) Limitations of galvanometer sensitivity and feed-resistor setting (0·05–0·5%).

The resultant uncertainty for powers between 20 and 200 mW is about ±1%, while for the range 20–2 mW it varies between ±1 and ±2·5%.

(5.5) Comparison with another Transfer Element

The film bolometer was compared with the thermistor mount mentioned in Section 3.3. This comparison was limited to a power of 2 mW and a frequency range 200 Mc/s–1 Gc/s because of the limitations of the thermistor mount for an accuracy of

±1%. Differences between the results obtained by the two milliwattmeters were of the order of 2%, which showed that the uncertainties were not greater than the estimated values.

(6) CONCLUSIONS

The bolometer has been found useful as an absolute r.f. milliwattmeter for the frequency band 200 Mc/s–4 Gc/s. The estimated uncertainty in its performance is ±1% for the range 20–200 mW, including the uncertainty arising from a moderate source mismatch. With power down to 2 mW, the uncertainty is increased.

The design is not restricted to any particular size; in fact, by scaling the linear dimensions in proportion to the square root of the maximum power, while retaining the 5 : 1 ratio of length to diameter, different power ranges with the same maximum to minimum power ratio can be covered with similar uncertainties. The same galvanometer sensitivity and the same limitations of ambient temperature variation apply to all sizes of element, but the rating of the standard d.c. bridge arm has to be increased proportionally as the maximum power is increased. With the larger sizes, the upper-frequency limit is reduced on account of the larger maximum diameter of the outer conductor; in fact, a power of 2 watts is probably the limit for a coaxial film bolometer usable up to 3 Gc/s. For the measurement of higher powers, a calibrated precision directional coupler or a dissipative attenuator can be used with the bolometer.

(7) REFERENCES

(1) BLEANEY, B.: 'Radio-Frequency Power Measurement by Bolometer Lamps at Centimetre Wavelengths', *Journal I.E.E.*, 1946, **93**, Part IIIA, p. 1378.

(2) GAINSBOROUGH, G. F.: 'Some Sources of Error in Microwave Milliwattmeters', *ibid.*, 1948, **95**, Part III, p. 229.

(3) CARLIN, H. J., and SUCHER, M.: 'Accuracy of Bolometric Power Measurements', *Proceedings of the Institute of Radio Engineers*, 1952, **40**, p. 1042.

(4) KERNS, D. M.: 'Determination of Efficiency of Microwave Bolometer Mounts from Impedance Data', *Journal of Research, National Bureau of Standards*, 1949, **42**, p. 579.

(5) GINZTON, E. L.: 'Microwave Measurements' (McGraw-Hill, 1957), Section 3.4.

(6) BECKER, J. A., GREEN, C. B., and PEARSON, G. L.: 'Properties and Uses of Thermistors', *Bell System Technical Journal*, 1947, **26**, p. 170.

(7) LANE, J. A.: 'Measurements of Efficiency of Bolometer and Thermistor Mounts by Impedance Methods', *Proceedings I.E.E.*, Paper No. 2318 R, September, 1957 (**104** B, p. 485).

(8) STREET, R.: 'The Absolute Measurement of Low Power at 3 000 Mc/s', *ibid.*, Paper No. 838 M, May, 1949 (**96**, Part III, p. 237).

(9) BAILEY, R. A., FRENCH, H. A., and LANE, J. A.: 'The Comparison and Calibration of Power-Measuring Equipment at Wavelengths of 3 cm and 10 cm', *ibid.*, Paper No. 1689 R, September, 1954 (**101**, Part III, p. 325).

(10) LANE, J. A.: 'Transverse Film Bolometers for the Measurement of Power in Rectangular Waveguides', *ibid.*, Paper No. 2488 R, January, 1958 (**105** B, p. 77).

(11) HARRIS, I. A.: 'The Theory and Design of Coaxial Resistor Mounts for the Frequency Band 0–4 000 Mc/s', *ibid.*, Monograph No. 132 R, May, 1955 (**103** C, p. 1).

(12) WOODS, D.: 'A Coaxial Connector System for Precision R.F. Measuring Instruments and Standards' (to be published).

621.375.4 : 621–526

The Institution of Electrical Engineers
Paper No. 3134 M
Jan. 1960
©

A TRANSISTOR QUADRATURE SUPPRESSOR FOR A.C. SERVO SYSTEMS

By I. C. HUTCHEON, M.A., A.M.I.Mech.E., Associate Member, and D. N. HARRISON.

(*The paper was first received* 31*st March, in revised form* 5*th May, and in final form,* 3*rd July,* 1959.)

SUMMARY

If the response angle of the demodulator in an a.c. servo system is not zero, any phase error in the input carrier signal causes an error in the servo output. Correcting the response angle to zero eliminates the output error, but is often impracticable and does not significantly reduce the residual signal at balance. This signal may saturate the amplifier, so making the servo insensitive or even completely inoperative.

Both effects are eliminated if a quadrature signal of appropriate magnitude is fed to the amplifier input so as to suppress the residual signal. If the phase error and hence the residual signal are not stable, the suppression must be controlled automatically, e.g. by a resistive element whose gain is varied by the residual signal itself.

The paper describes a quadrature suppressor which uses four low-power transistors and three indirectly heated thermistors. Two of the transistors form a pre-amplifier, and two operate in switched mode as a demodulator, driving two of the thermistor heaters differentially. The corresponding thermistor beads act as a variable voltage divider which is used to control a suppression signal of precise quadrature phase. The transistors are all direct coupled, and the third thermistor provides d.c. feedback around the suppressor giving thermal stability and constant output resistance characteristics in ambient temperatures between 0 and 60° C. The device is analogous to a servo-driven slide-wire of moderate linearity, and will operate with a loop gain of several hundred without instability.

LIST OF PRINCIPAL SYMBOLS

β = Gain of quadrature loop.
I = Net steady direct current applied to demodulator, amp.
I_b = Peak value of full-wave rectified bias current applied to demodulator, amp.
I_s = Peak value of a.c. signal applied to demodulator, amp.
I_{ms1} = Mean-square current in TH_1 heater, (amp)^2.
I_{ms2} = Mean-square current in TH_2 heater, (amp)^2.
R_L = Demodulator load resistance, ohms.
V_1 = Reference-phase negative feedback signal, volts.
V_2 = Quadrature-phase negative feedback signal, volts.
V_q = Quadrature component of input signal, volts.
V_r = Reference-phase component of input signal, volts.
V_s = Peak value of demodulator alternating supply voltage, volts.
α_1 = Small phase shift applied to V_1, deg.
α_2 = Small phase error in V_2 (from 90°), deg.
θ_1, θ_2 = Demodulator conducting angles, alternate half-cycles, deg.
ϕ = Response angle of servo amplifier and demodulator, deg.
ω = Angular frequency of signal, rad/s.

(1) INTRODUCTION

Two conditions must be satisfied for an a.c. servo system to achieve a perfect balance, namely equality in both magnitude and phase of the input and the feedback carrier signals. D.C. servo systems, by contrast, require merely equality in magnitude. If the two signals differ in phase by even one or two degrees, quite substantial errors may arise in the magnitude of the feedback signal and hence in the servo output if the response angle of the demodulator is not zero. Furthermore, the residual error signal which exists when the system has reached balance may saturate the amplifier, so causing a loss of sensitivity which may be partial or complete.

The output error can be removed by correcting the response angle, but this is frequently a troublesome procedure and has little effect on the size of the residual signal. It is much more satisfactory to suppress the quadrature effects where they arise by injecting a quadrature signal of opposite phase at the a.c. amplifier input. If the quadrature effects are stable the compensating signal may be manually adjusted, but if they are variable some form of automatic compensation becomes necessary.

(2) QUADRATURE EFFECTS IN A.C. SERVO SYSTEMS

(2.1) Causes of Quadrature

Quadrature effects may be caused either by the injection of a signal which is not of reference phase, as, for example, a pick-up signal, or by phase shifts in the input or the feedback circuit. The rotation of a signal by a few degrees has little effect on its magnitude, but introduces a component in quadrature with it of 1·75% per degree of rotation. Phase shifts are commonly introduced by coupling transformers and by the combined effect of source resistance and transmission-line self-capacitance.

Good system design reduces these causes to a minimum by the use of high-quality transformers, low-capacitance low-inductance transmission lines, adequate screening of high-impedance points and so on. Nevertheless it is frequently impossible to eliminate them altogether, and in many cases the residual quadrature effects are not constant.

To take an example, the output signal from a small induction-type liquid flowmeter is typically a few millivolts at mains frequency. The source resistance depends on the electrical conductivity of the liquid and varies from a few hundred ohms or less to as high a value as the measuring system will tolerate, in one case about 40 kilohms. If the output from the flowmeter is transmitted to a self-balancing a.c. recorder via a 100 ft cable whose self-capacitance is 30 pF/ft, a quadrature component is introduced which varies, in this case, from almost zero to about 4% of the transmitted signal. In addition, the flowmeter output may contain a relatively stable quadrature component owing to induction effects in residual circuit loops.

Worse effects arise in systems for measuring the electrical conductivity of liquids. Capacitance effects at the electrode surfaces, due to the presence of thin polarizing gas films, give rise to quadrature components which may be 10% or more in magnitude and which vary with the liquid conductivity, the state of the surface and other parameters.

(2.2) Effects on the System

Consider the a.c. servo system shown diagrammatically in Fig. 1(*a*), which might, for example, be a self-balancing industrial

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
Mr. Hutcheon and Mr. Harrison are with George Kent Ltd.

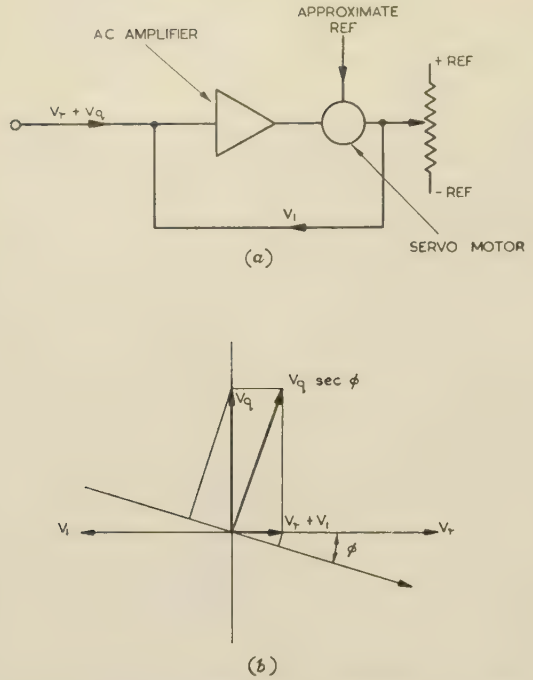


Fig. 1.—A.C. servo system without quadrature suppression.
(*a*) Block diagram.
(*b*) Vector diagram of signals at amplifier input.

recorder. The motor demodulates the amplified carrier signal, and drives a slide-wire contact, so providing a negative feedback voltage, V_1, of reference phase which reduces the net input towards zero. Fig. 1(*b*) is a vector diagram of the signals applied to the amplifier input.

The motor has its maximum response to input signals at some angle ϕ, which in general is not zero, and zero response to signals at $\phi \pm 90°$. When the system is in a state of balance, the resultant input signal at angle ϕ is too small to turn the motor and may be regarded as zero if the loop gain is high. Thus

$$(V_r + V_1)\cos\phi - V_q \sin\phi = 0 \quad . \quad . \quad . \quad (1)$$

and the feedback signal is in error by a fractional amount

$$\frac{V_r + V_1}{V_1} = \frac{V_q}{V_1}\tan\phi \simeq \frac{V_q}{V_r}\tan\phi \quad . \quad . \quad . \quad (2)$$

There exists also a residual error signal at $\phi \pm 90°$ whose magnitude is

$$[(V_r + V_1)^2 + V_q^2]^{1/2} = V_q \sec\phi \quad . \quad . \quad . \quad (3)$$

The response angle, ϕ, should desirably be zero but often is not very closely controlled. Production spreads in values of the phase-balancing and resonating capacitors associated with a 2-phase induction motor, for example, may cause phase errors of 10° or 20°, whilst variations in motor inductance and phase shift through the amplifier may increase this figure. If $\phi = 20°$, then $\tan\phi = 0{\cdot}36$, and a 4% quadrature component produces an error in output of 1·44%.

Saturation of the amplifier by the residual signal may also be a serious effect. A 1% quadrature signal applied to the input of a typical self-balancing recorder, adjusted to be sensitive to input signals equal to 0·1% of the span, just produces a limiting signal at the amplifier output. This in itself does not seriously impair the performance, but any further increase progressively slows the response to in-phase signals until, with about 10% quadrature component applied, the motor fails to turn.

(3) METHODS OF QUADRATURE SUPPRESSION

Reducing the response angle to zero is not a good solution to the problem for two reasons. First, it does not significantly reduce the residual signal $V_q \sec\phi$, and the possibility of saturation remains. Secondly, it is often troublesome, since the amplifier and motor or other demodulator are usually required to be replaceable units, and each must be phase corrected individually. It is therefore better to eliminate the quadrature signal itself.

(3.1) Manual Feedback Suppression

When the quadrature effects are stable, they can be cancelled by the injection of an anti-phase signal which is adjusted manually until the residual signal at balance is a minimum. The adjustment must be made after the system has been installed, and usually requires some form of detector to be available to permit observation of the residual signal. Alternatively, a 90° shift may be introduced temporarily into the amplifier or the demodulator, and the demodulator output itself, e.g. motor rotation, used as the indication. This method of quadrature suppression is the simplest, but is of no value if the quadrature effects are variable.

(3.2) Automatic Series Suppression

Various methods[1] have been used to provide continuous automatic elimination of the quadrature signal, and all can be classified either as series or feedback devices.

Series quadrature eliminators connect in series with the input signal and pass only the signal component which is of reference phase. Some operate by demodulating the signal, smoothing the resultant output and remodulating at reference phase; others[2] operate by a sampling process which gives a faster response. If the signals are small it is difficult to reduce drift effects to a suitably low level, and resort has sometimes been made[3] to the use of mechanical choppers. Apart from the difficulty of obtaining a known and stable reference phase by this method it appears to be an undesirable complication. Since the input signal has to pass through the device, its response must be fast, and this introduces further design problems.

(3.3) Automatic Feedback Suppression

By far the most satisfactory method is the use of automatically controlled feedback suppression. As with manual suppression, an anti-phase quadrature signal is fed to the input of the servo system, but in this case it is controlled automatically by the residual signal itself. In order to prevent further error, the feedback signal must contain no reference-phase component, and it is usual to demodulate the amplified residual signal and use it to control a feedback signal whose phase can be determined with precision. The quadrature feedback loop has no effect on the response of the main servo system, and can be quite slow acting, since quadrature effects usually vary slowly. The signal levels are much higher than in the series case, and drift effects are not encountered. Quadrature effects are reduced by a factor equal to the gain of the quadrature loop, and it is seldom necessary to make this exceed about 100. The theory of feedback quadrature suppression is outlined in Section 7.1.

One method is to use a subsidiary servo motor, which is driven directly from the main servo amplifier, but has its power winding supplied so that it responds to quadrature signals. The motor drives a slide-wire which is connected across a supply of precise quadrature phase, and so provides the necessary feedback.

Servo motors and slide-wires are not cheap components, however, and the system which will be described replaces them by a pair of indirectly heated thermistors. A subsidiary transistor

amplifier is used to drive them, so making the suppressor sufficiently sensitive for use with virtually any servo system.

The final device is basically a resistive voltage divider whose gain is controlled in a moderately linear fashion between $\pm 75\%$ by an a.c. input of $\pm 1\,\mu A$.

(4) DESIGN OF TRANSISTOR QUADRATURE SUPPRESSOR

(4.1) Basic Principles

Fig. 2(*a*) is the block diagram of a quadrature suppressor employing two indirectly heated thermistors to control the suppression signal. Fig. 2(*b*) shows an alternative output arrangement, and Fig. 3 is the diagram of a circuit which uses four low-power germanium transistors for pre-amplification and demodulation.

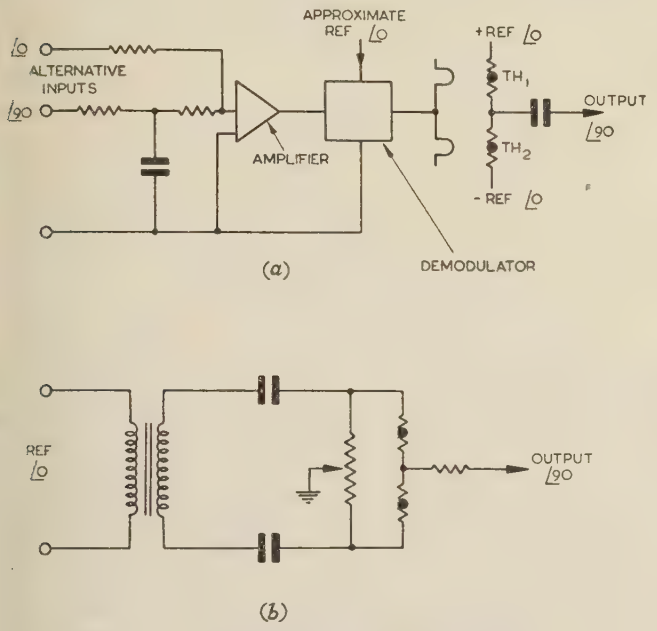


Fig. 2.—Block diagram of quadrature suppressor.
(*a*) With simplest output circuit.
(*b*) Alternative output circuit.

The output signal of the servo amplifier is applied to the transistor amplifier and demodulator via either a resistor or an *RCR* circuit giving a lag of approximately 90°. Which of these is chosen depends on the phase of the servo carrier signal, the aim being to arrange that the undesired quadrature effects drive the demodulator, which, for convenience, operates directly from the supply.

The demodulator controls the power supplied to the heaters of two thermistors TH_1 and TH_2, any signal causing an increase in the dissipation of one and a decrease in that of the other. This lowers the resistance of one thermistor bead and raises that of the other, so unbalancing the voltage divider formed by the two beads. The demodulator is phase sensitive, and a reversal of signal phase reverses the direction of unbalance. Although the heaters are supplied with alternating current, thermal lag in the beads provides a smoothing effect and prevents modulation of the bead resistances at carrier frequency.

The voltage divider may be connected across a push-pull a.c. supply of reference phase, as shown in Fig. 2(*a*), the suppression current then being derived via a small capacitor whose reactance is chosen to be much higher than the divider output resistance plus the final load. This ensures that the suppression current is in quadrature with the reference to any desired degree of accuracy. The power which may be dissipated in the thermistor beads is limited to a few milliwatts, however, and larger outputs can be obtained if the 90° phase shift is derived before the voltage divider, as shown in Fig. 2(*b*), instead of after it. In many cases the simpler arrangement is adequate. Either circuit may be provided with a zero adjustment, as shown in Fig. 2(*b*), to permit compensation for initial unbalance in the bead resistances. Again, however, this is usually unnecessary, since the quadrature signal due to the unbalance acts through the whole quadrature loop and readjusts the heater dissipations until it is itself suppressed by a factor equal to the loop gain. The only disadvan-

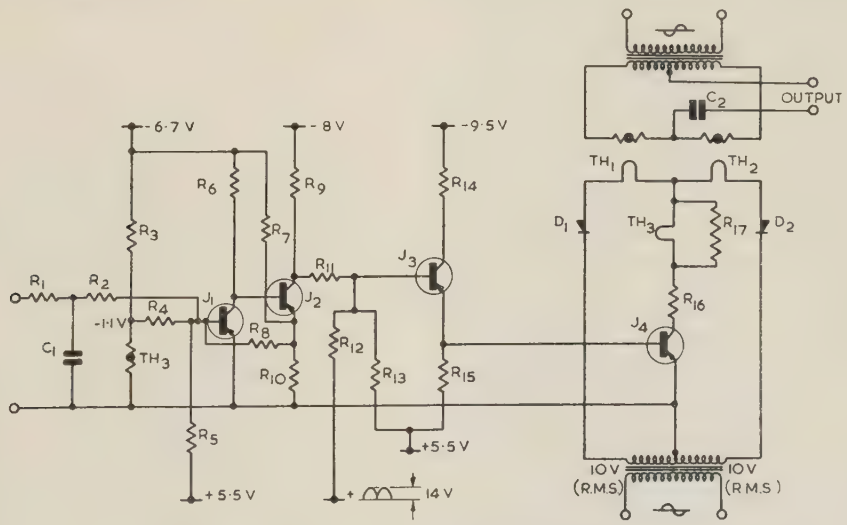


Fig. 3.—Circuit diagram of quadrature suppressor.

R_1	1 MΩ
R_2	56 kΩ
R_3	5·6 kΩ
R_4	6·8 kΩ
R_5	47 kΩ
R_6	15 kΩ
R_7	10 kΩ
R_8	6·8 kΩ
R_9	2·7 kΩ
R_{10}	100 Ω
R_{11}	2·2 kΩ
R_{12}	27 kΩ
R_{13}	10 kΩ
R_{14}	4·7 kΩ
R_{15}	10 kΩ
R_{16}	220 Ω
R_{17}	240 Ω

TH_1 bead, 1·3 kΩ nominal.
TH_2 bead, 1·3 kΩ nominal.
TH_3 bead, 1·2 kΩ nominal.

C_1 0·5 μF
C_2 0·03 μF

D_1, D_2, type OA5

J_1, type OC45; J_2, J_3, J_4, type OC76

tage is that part of the available swing of the voltage divider in one direction is thereby used up.

It would be possible, in cases where the servo amplifier had adequate output power, to drive the thermistors directly from this output and a quadrature power supply. The inclusion of an additional amplifier, however, has two significant advantages. First, it reduces the loading on the servo amplifier to a negligible level and so permits the suppressor to be used with virtually any servo system. Secondly, it allows compensation of the otherwise considerable effects of ambient temperature on the thermistor characteristics.

An uncompensated thermistor voltage divider must operate with relatively low heater dissipations so as to allow safe working at the highest ambient temperature which may be encountered, and at the same time to provide an adequate margin for changes in dissipation with driving signal. Under these conditions, the output resistance and the resistance between the ends of the divider vary widely with normal changes in ambient temperature, both effects making it difficult to design suitably accurate phase-shifting circuits.

The difficulty is overcome by the use of a third thermistor, which controls a direct feedback current around the complete quadrature suppressor in such a manner that the power supplied to the output thermistors is reduced as the ambient temperature rises. This keeps the output characteristics substantially independent of the ambient temperature between 0 and 60° C. It also stabilizes all the transistors against temperature effects up to about 70° C in a manner proposed by Candy,[4, 5] and obviates the need for a decoupling capacitor. Thermal lag in the thermistor provides the necessary smoothing.

(4.2) Circuit Description

Transistors J_1 and J_2 in Fig. 3 constitute the a.c. amplifier,[4] whose current gain is determined close to the value R_8/R_{10} by negative feedback from the emitter of J_2 to the base of J_1. The output current of J_2 divides about equally between R_9 and R_{11}, giving a current gain from the base of J_1 to the base of J_3 which is approximately equal to $\frac{1}{2}R_8/R_{10} = 34$.

The demodulator comprises transistors J_3 and J_4, diodes D_1 and D_2, and the centre-tapped a.c. supply. J_3 and J_4 operate together and can be regarded as a single transistor whose current gain exceeds 1 000, the demodulator equivalent circuit thus being as shown in Fig. 4. A positive full-wave-rectified bias current is supplied to the base of J_3, which combines with the net steady negative bias, controlled by J_2, to switch the J_3–J_4 combination on or off according as the resultant signal is negative or positive. In the absence of any a.c. drive from the amplifier, thermistors TH_1 and TH_2 each receive equal heater currents comprising two pulses every other half-cycle. Any supply-phase signal applied to the demodulator modifies the conducting angles, unbalances the heater dissipations, and so controls the gain of the voltage divider between absolute limits of $\pm 100\%$, which are never quite achieved in practice. The operation of the demodulator is analysed fully in Section 7.2, and the relevant waveforms are shown in Fig. 5(*a*)–(*e*). With the circuit parameters given in Fig. 3, a current of $\pm 32\,\mu$A (r.m.s.), applied from J_2 to the demodulator, swings the voltage-divider output by $\pm 75\%$ of its theoretical maximum value. Within these limits, transistors J_3 and J_4 switch sharply, and the overall demodulator characteristic is fairly linear, as can be seen from Fig. 6.

The overall characteristic of the quadrature suppressor, from the base of J_1 to the output of the voltage divider, is therefore approximately equal to $\pm 75\%$ per μA (r.m.s.). The optional *RCR* input network has a gain of $1/\omega C_1 R_1 R_2$, which, at 50 c/s and with the specified component values, is equal to $0{\cdot}1\,\mu$A/volt. The overall gain of the quadrature suppressor preceded by this

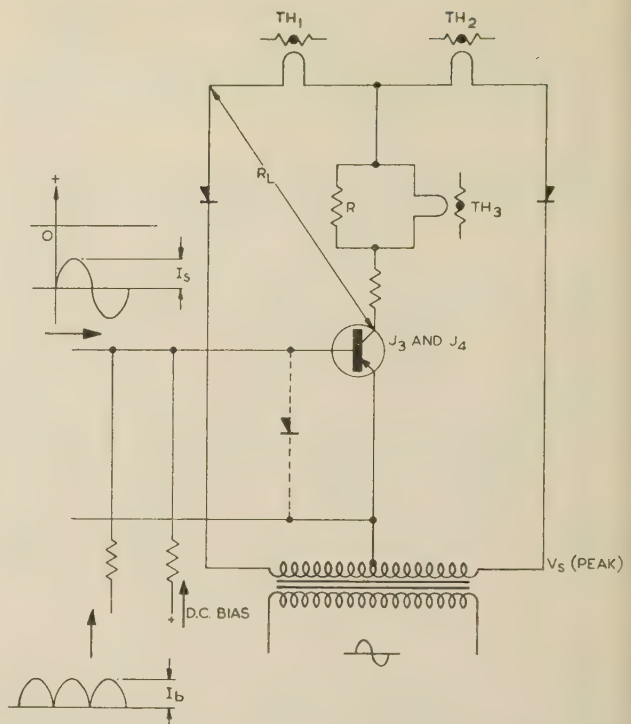


Fig. 4.—Demodulator equivalent circuit.

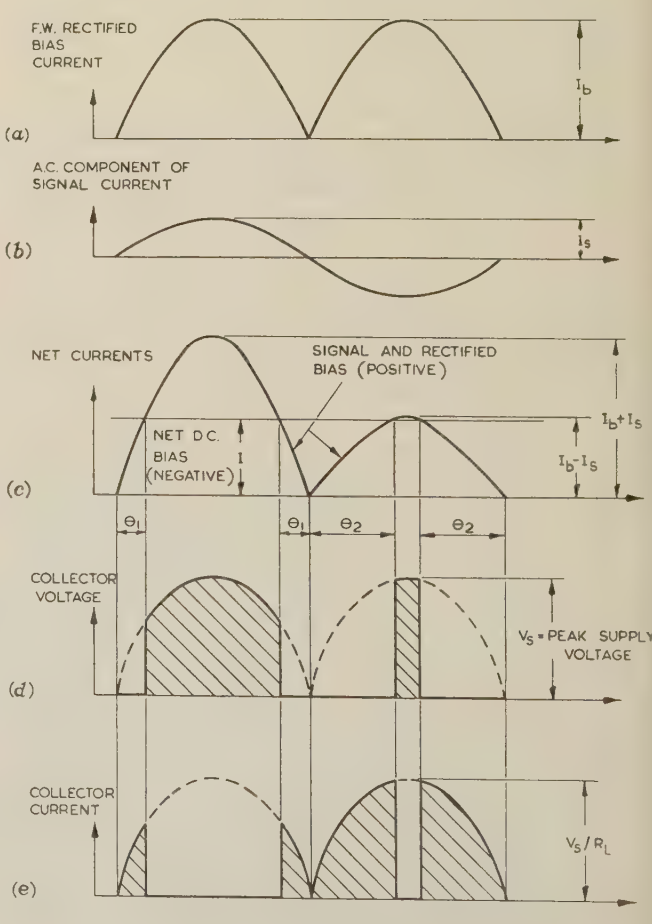


Fig. 5.—Demodulator waveforms.

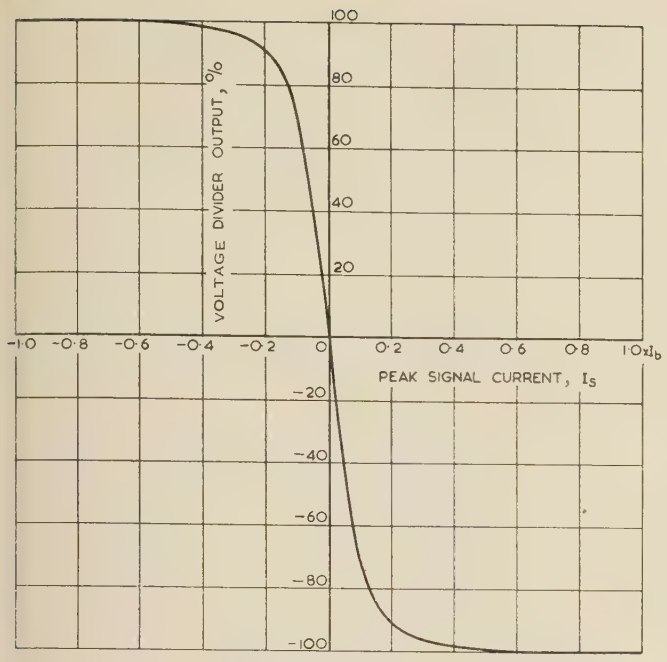


Fig. 6.—Gain of demodulator and voltage divider at 30° C.

network is therefore approximately ±75% per 10 volts (r.m.s.) at 50 c/s.

The amplifier and demodulator contain no reactive components, and the upper limit of operating frequency is determined by the ability of J_3 and J_4 to switch rapidly, being typically a few kilocycles per second for low-frequency transistors. At low supply frequencies, the bead resistances start to follow the pulses of the heater current, and the lower limit of operating frequency is determined by the thermal lag between the heaters and the beads. With typical thermistors, bead-resistance modulation is negligible for carrier frequencies above 30 c/s.

The a.c. feedback applied to the base of J_1 gives the amplifier a low input impedance, the typical value being about 150 ohms. This is useful in that it permits the input current to be determined quite accurately by a resistance of a few kilohms, and the suppressor to be driven fully by a supply-phase signal of a few millivolts. Response of the suppressor is limited by the output thermistors, and a large change in output takes several seconds to complete, as can be seen from the curves in Fig. 7. This is seldom a disadvantage, however, and has the advantage that the gain of the complete quadrature-suppressing loop may be increased to about 500 before instability occurs. This provides a useful safety margin, since loop gains of 50 or 100 are usually quite adequate to render quadrature effects negligible.

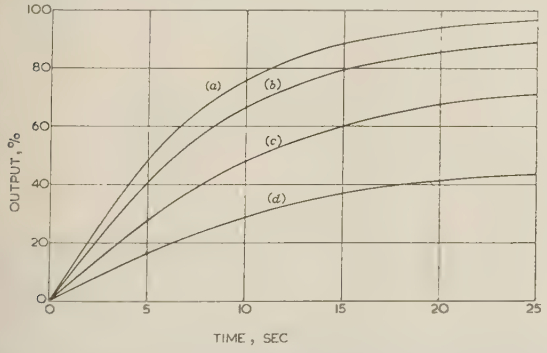


Fig. 7.—Response of quadrature suppressor to step-input changes.
(*a*) Input 0–30 volts (r.m.s.).
(*b*) Input 0–20 volts (r.m.s.).
(*c*) Input 0–10 volts (r.m.s.).
(*d*) Input 0–5 volts (r.m.s.).

(4.3) D.C. Stabilization

The entire circuit is direct coupled, and stabilization against ambient temperature changes is effected by overall d.c. feedback applied via the third thermistor, TH_3, whose thermal lag prevents any feedback of the a.c. signal. Inter-stage bias arrangements are described in Section 7.4.

The bead of thermistor TH_3 is connected in the base circuit of J_1 in such a manner that a change of ±15% in its nominal resistance varies the base current by about ±20 μA. This is sufficient to counter the effects of varying collector leakage currents in J_1 (0–30 μA maximum between 0 and 60° C for a type OC45 transistor) and of initial differences in base current due to the spread in gain between transistors. Drifts which occur after J_1 can be largely neglected, since their effects, when referred to the input, are divided by the gain of the previous stages.

A further ±7% change in the resistance of the TH_3 bead would be sufficient to change the direct current in R_{11} by ±0·35 mA, so taking J_3 and J_4 from the always-on to the always-off condition and varying the dissipation in the TH_3 heater between 0 and 30 mW. This in turn would vary the resistance of the TH_3 bead very widely. Thus a ±22% change in the resistance of the TH_3 bead is sufficient to counter all ambient temperature effects and, at the same time, to provide a much greater compensating change in resistance via the feedback loop. The bead resistance is therefore controlled within approximately ±20% of the nominal value for all ambient temperatures between 0 and 60° C.

The nominal resistance of the TH_3 bead is chosen as 1·2 kilohms, which corresponds to a dissipation of 16 mW at 30° C. Since about 1 mW is dissipated in the bead itself, the heater receives 15 mW at 30° C. The heater currents of TH_1 and TH_2 flow alternately through the heater of TH_3 which is shunted so that, when there is no drive, all three heaters receive the same power whatever the ambient temperature. The method of calculating the shunt resistance is given in Section 7.3. The bead resistances of TH_1 and TH_2 are thus controlled within about ±20% of 1·3 kilohms (which is the value corresponding to 15 mW and 30° C), for ambient temperatures between 0 and 60° C. In practice, the limits are somewhat wider, owing to the spread in the thermistor nominal characteristics, but the final result is a very useful stabilization of the output-resistance characteristics against temperature variations.

(4.4) Characteristics of Voltage Divider

When drive is applied to the demodulator, the heater dissipations of the two output thermistors increase and decrease respectively by equal amounts about a value of 15 mW at 30° C, until one is 30 mW and the other is zero.

At higher ambient temperatures the overall d.c. feedback enforces a fall in the dissipation with zero drive of 2 mW for every 10° C rise, and the output swing is somewhat restricted, since one heater becomes fully cut off before the other can be fully driven. At lower ambient temperatures there is a similar restriction, since one heater receives the maximum dissipation of 30 mW before the other is cut off. These effects are shown in Table 1 for design-centre and limit values of ambient temperature.

In even the extreme cases, a swing of ±9 mW is available for each heater, and allows the voltage divider to provide considerably more than ±75% of the theoretical maximum output.

Table 1

Ambient temperature	Heater dissipations		
	TH_1 or TH_2, full drive	TH_1, TH_2 and TH_3, no drive	TH_2 or TH_1, full drive
deg C	mW	mW	mW
0	12	21	30*
30	0*	15	30*
60	0*	9	18

* Limiting factors.

The drive required is slightly greater than that needed at 30° C, but this effect is not significant.

When driven sufficiently to provide a 75% output, the beads of TH_1 and TH_2 have resistances of about 600 ohms and 3·8 kilohms, while each bead has a resistance of about 1·3 kilohms when there is no drive. The output resistance of the voltage divider thus varies from about 700 to 500 ohms as the drive is increased, while the resistance between the two ends of the divider varies between about 2·5 and 4·5 kilohms. These limits are increased somewhat by the initial tolerance on bead resistance and by the ±20% variation required by the d.c. stabilizing loop. It is worthy of note that, so long as the feedback operates, neither supply-voltage variations nor exceptionally high ambient temperatures can overload the thermistors, although some care is necessary in applying a signal to the beads. Should the feedback fail to operate, and the supply voltage increase by 10%, the maximum power which can be supplied is 36 mW, and no damage can occur if the ambient temperature does not exceed 40° C.

(4.5) Applications

Almost any servo amplifier will supply enough power to drive the quadrature suppressor, but some applications may require use of the output circuit arrangement shown in Fig. 2(*b*) in order that adequate feedback power is available. A supply of about 1–0–1 volt (r.m.s.) may be connected across the voltage divider without overloading the thermistors, and gives an output of at least ±0·75 volt from a source resistance of 500–700 ohms approximately.

The arrangements of Fig. 2(*a*) and (*b*) both advance the reference by 90° and exaggerate any harmonics that are present. Even so, the residual harmonic signal in the servo amplifier, when the suppressor is operating, is many times smaller than the fundamental quadrature signal which is being suppressed, and usually causes no trouble. Some cancellation of harmonics may, in fact, occur if any are present in the quadrature signal which exists initially. Obtaining an output of useful magnitude which lags the reference by 90° is more difficult, since the divider is resistive and may require the use of further amplification.

Quadrature effects are reduced by a factor equal to the gain of the complete quadrature loop, and values below 100 are usually adequate. The upper limit is determined by the onset of instability, and is typically about 500. The quadrature suppression signal should be applied in such a way that variations in source impedance of the servo input signal do not require the suppressor output to vary. Thus the feedback should be arranged to develop a voltage in series with the input to the servo amplifier if the quadrature effect is a series voltage, or applied as a current if the quadrature effect is generated in this way.

Two units connected as shown in Fig. 8 suppress the reference-phase and quadrature components, respectively, of the input

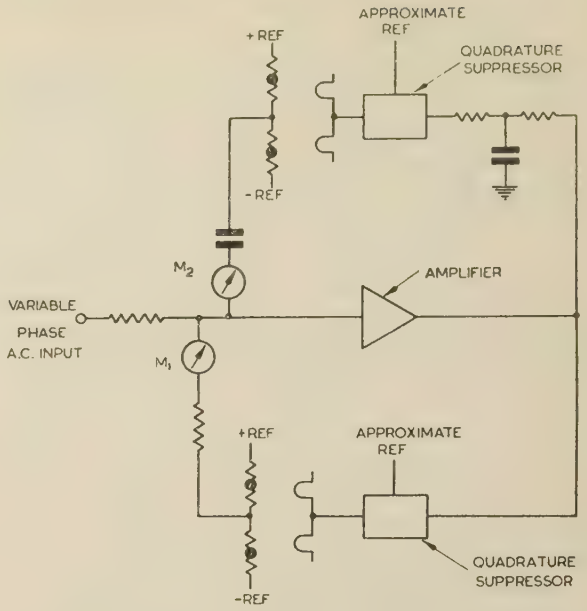


Fig. 8.—Use of two units for phase measurement.

current. Meters M_1 and M_2, indicating the values of the two components, constitute a phasemeter whose accuracy is limited mainly by that of the meters. Moderate phase shifts through the subsidiary amplifier are of no importance.

It is also possible to make use of the slow response of the voltage divider in time-sharing circuits. For example, the divider can be shared between two circuits to perform multiplication or division.[6] In one circuit its gain is set equal to the ratio V_1/V_2 of two direct voltages, while in the other this gain is used to control the ratio V_3/V_4 of two other voltages. Thus $V_1V_4 = V_2V_3$ and the system multiplies or divides in a slow but accurate fashion. Two single-pole double-throw synchronous switches and an additional a.c. amplifier and demodulator are required. The whole system operates on a.c. signals, and there are no problems of d.c. drift.

Other interesting possibilities can be envisaged which are an extension of the above and depend on the use of rapid multiway switching and a number of analogue memory (smoothing) circuits.[7] These include the generation of power series by repeated multiplication, and the multiplication of many inputs by the same factor.

Frequency sharing is also feasible, it being possible, for example, to set up a direct voltage whose ratio to another is equal to the ratio of two mains-frequency a.c. voltages. The output of a differential transformer can be converted into a direct voltage or current in this manner.[8]

(5) ACKNOWLEDGMENTS

The Directors of George Kent Ltd. are thanked for giving permission to publish the paper.

(6) REFERENCES

(1) Control Engineers Handbook (McGraw-Hill, 1958), Section 6.18, p. 77.

(2) Fennick, B.: 'Phase-Selective Gate Rejects Quadrature', *Electronics*, 1958, **31**, p. 92.

(3) Williams, A. J., and Payne, J. F.: 'A.C. Null-Type Recorder with Balancing Amplifier which provides Damping and suppresses the Quadrature Component', *Transactions of the American I.E.E.*, 1953, **72**, p. 611.

(4) CHAPLIN, G. B. B., and OWENS, A. R.: 'A Transistor High-Gain Chopper-Type D.C. Amplifier', *Proceedings I.E.E.*, Paper No. 2442 M, November, 1957 (**105** B, p. 258).
(5) CHAPLIN, G. B. B., CANDY, C. J. N., and COLE, A. J.: 'Transistor Stages for Wide-Band Amplifiers', *ibid.*, Paper No. 2892 E, May, 1959 (**106** B, Suppl. 16, p. 760).
(6) HUTCHEON, I. C.: 'The Design and Application of a Synchronouse Convertor: Part II', *British Communications and Electronics*, 1958, **5**, p. 602.
(7) HUTCHEON, I. C.: 'An Iterative Analogue Computer for use with Resistance Network Analogues', *British Journal of Applied Physics*, 1957, **8**, p. 370.
(8) British Patent Application No. 32483, 1957.
(9) British Patent Application No. 13560, 1959.

(7) APPENDICES

(7.1) Theory of Feedback Quadrature Suppression

Consider Fig. 9, which is a block diagram of an a.c. servo system with quadrature suppression applied via a second feedback loop. The amplifier and motor drive a slide-wire contact

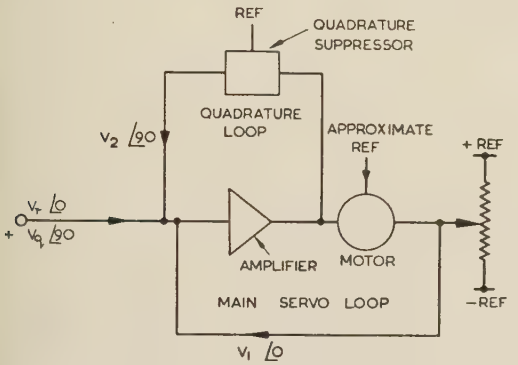


Fig. 9.—A.C. servo system with feedback quadrature suppression.

which provides the main negative feedback signal, V_1, at reference phase. V_1 is required to be equal in magnitude to the reference-phase component, V_r, of the input signal, in spite of the presence of a quadrature input component, V_q. In the absence of any suppression, as was shown in Section 2.2, there would exist a substantial error equal to $(V_q/V_r)\tan\phi$, where ϕ is the response angle of the amplifier and motor. Accurate operation is made possible by the quadrature feedback signal, V_2, which reduces the effect of V_q to small proportions.

If both feedback loops are assumed to have infinite gain, and all the signals have precisely the phase angles shown in Fig. 9, the two loops are entirely non-interacting, and, since the net input to the amplifier must be zero, $V_2 = -V_q$ and $V_1 = -V_r$, which is the desired relationship. This holds good in theory for any value of ϕ, but in practice, for any value of ϕ which permits the two loops to retain a reasonable fraction of their gain.

In practical systems, high loop gains can usually be achieved, but coupling transformers, etc., frequently cause a small phase error in the servo feedback signal, V_1, while it is never possible to generate a quadrature feedback signal precisely at 90° to the reference. Input phase errors also exist, and are already implied in the assumption that V_q exists. These phase errors, taken together, give rise to a small residual error in the servo output whose value we shall calculate. The following small phase errors, which for simplicity are regarded as phase advances, are assumed to be present: α_1, acting on V_1, and α_2 acting on V_2. V_r is assumed to be 90° in advance of the reference.

Rotation of any a.c. signal, V, through a small angle, α, has only a very small effect on its magnitude (e.g. 0·06% change for a 2° rotation), and this will be neglected since α_1 and α_2 are in fact small. The same rotation, however, introduces a quadrature component of magnitude αV, whose fractional value is therefore 1/57 (i.e. 1·75%) per degree. This effect is not negligible, being in fact the cause of the error in output. No assumption is made as to the magnitude of V_q, since this is sometimes relatively large.

Both feedback loops are provisionally assumed to have high gains, and the resultant signal applied to the input of the servo amplifier is therefore zero. Consequently, the sum of the components of V_r, V_q, V_1 and V_2, when resolved in any direction, must be zero, Thus

$$\text{at } 0° \qquad V_r + V_1 - \alpha_2 V_2 = 0 \quad . \quad . \quad . \quad . \quad . \quad (4)$$

$$\text{and at } 90° \qquad V_q + \alpha_1 V_1 + V_2 = 0 \quad . \quad . \quad . \quad . \quad . \quad (5)$$

Eliminating V_2, which is unknown, from these equations,

$$V_1 + V_r + \alpha_2(V_q + \alpha_1 V_1) = 0 \quad . \quad . \quad . \quad . \quad (6)$$

whence the fractional error in V_1 is

$$\frac{V_1 + V_r}{V_1} \simeq -\alpha_2\left(\frac{V_q}{V_r} + \alpha_1\right) \quad . \quad . \quad . \quad . \quad (7)$$

V_1 and V_r representing signals of opposite phase which are very nearly equal in magnitude.

Since the error is proportional to α_2, it is clearly desirable to phase the quadrature feedback signal as accurately as possible. If α_2 is held below 0·5°, which is a reasonable aim, the output error due to a 10% quadrature input component is always less than 0·1%.

It is worthy of note that α_1 appears in eqn. (7) as a second-order effect, being in fact precisely analogous to the rotation V_q/V_r assumed for the input signal. Thus variations in phase of the servo feedback signal to within a few degrees have very little effect on the servo accuracy when quadrature suppression is applied.

It has so far been assumed that the gain of the quadrature loop is infinite. In fact it is not, and we require to know what value of loop gain is necessary to justify the assumption and keep the error within specified limits. A similar analysis, in which the quadrature loop gain is assumed to have a finite value β, gives the following expression for the fractional error in servo output:

$$\frac{V_r + V_1}{V_1} = \left(\frac{1}{\beta}\tan\phi - \alpha_2\right)\left(\frac{V_q}{V_r} + \alpha_1\right) \quad . \quad . \quad (8)$$

where ϕ is the response angle of the demodulator. Both terms in the first bracket may be either positive or negative, and so may add, giving an error which may be twice the lowest possible error (for $\beta = \infty$) when

$$\beta_{crit} = \frac{1}{\alpha_2}\tan\phi \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

If β is in fact made a few times greater than β_{crit}, the error closely approaches the lowest possible value determined by α_2. No advantage is gained by making it very many times greater, and if α_2 is very small there may be no need to approach the limit. If $\alpha_2 = 0{\cdot}5° = 1/114$ rad, for example, and $\phi = 20°$, then $\beta_{crit} = 40$, and if β is in fact made equal to 100, the error is

$$\left(\frac{0{\cdot}36}{100} \pm \frac{1}{114}\right)\left(\frac{V_q}{V_r} + \alpha_1\right)$$

which is within 1·4 times the value for infinite β and, in any

case, small. A loop gain of 100 is adequate for most purposes, and can readily be obtained.

The residual signal, and hence the probability of saturation, also reduces in proportion to $1/\beta$. There is no theoretical limit in this case, except that set by the finite gain of the main servo loop. In practice, however, when any reasonable degree of quadrature suppression has been applied, harmonic effects are the main factor tending to cause saturation, and these depend mainly on V_q and very little on β.

(7.2) Analysis of Demodulator

The basic demodulator circuit is given in Fig. 4, and the relevant waveforms appear in Figs. 5(*a*)–(*e*). The heater currents controlled by transistor J_4 never exceed 35 mA, and the J_3–J_4 combination, whose current gain is at least 1 000, is switched on and off by a change of 35 μA or less in the base current of J_3. This is very small in comparison with the signals applied to the J_3 base and can therefore be neglected.

Other assumptions are that the diode reverse leakage current is negligible, and that the forward voltage drop across J_4 and either diode is small in comparison with the demodulator supply voltage. The small drop which occurs in practice is compensated by reducing the load R_L from the theoretical value of 410 ohms to 390 ohms.

Four currents are supplied to the base of transistor J_3, namely:

A full-wave rectified positive bias current, whose peak value is I_b.

The a.c. signal, of variable magnitude, whose peak value is I_s.

A steady positive bias current of 0·6 mA.

A steady negative current drawn by J_2 and controlled by the overall d.c. feedback, which exceeds the bias of 0·6 mA, less any collector leakage in J_3, by a net amount I.

The rectified bias and signal waveforms, shown in Fig. 5(*a*) and (*b*), combine to give the waveform of Fig. 5(*c*), and this switches J_3 off or on according as it is greater or less than I. Each heater thus receives two current pulses every other half-cycle as shown in Fig. 5(*e*).

In practice, J_3 will not accept a positive base current, and its base potential makes a positive excursion whenever it is switched off. The circuit operation is unaffected, however, and description is simplified if the diode shown with dotted lines in Fig. 4 is assumed to exist. This carries the excess positive current, and the base of J_3 can be regarded as remaining at earth potential.

Switching occurs as the net signal applied to the J_3 base passes through zero, and the conducting angles θ_1 and θ_2, for alternate half-cycles, are given by

$$\sin\theta_1 = \frac{I}{I_b + I_s} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

and

$$\sin\theta_2 = \frac{I}{I_b - I_s} \quad . \quad . \quad . \quad . \quad . \quad (11)$$

whence

$$\frac{I_s}{I_b} = \frac{\sin\theta_2 - \sin\theta_1}{\sin\theta_2 + \sin\theta_1} \quad . \quad . \quad . \quad . \quad . \quad (12)$$

Whenever the transistors J_3 and J_4 are switched on, the instantaneous current in the load is equal to $V_s \sin\omega t/R_L$, whence the mean-square current in one heater is

$$I_{ms1} = \frac{\omega}{2\pi}\left(\frac{V_s}{R_L}\right)^2 \int_{\theta_1/\omega}^{(\pi-\theta_1)/\omega} (\sin\omega t)^2 dt$$

i.e.

$$I_{ms1} = \frac{1}{4}\left(\frac{V_s}{R_L}\right)^2 \frac{2\theta_1 - \sin 2\theta_1}{\pi} \quad . \quad . \quad . \quad (13)$$

while that in the other heater is

$$I_{ms2} = \frac{1}{4}\left(\frac{V_s}{R_L}\right)^2 \frac{2\theta_2 - \sin 2\theta_2}{\pi} \quad . \quad . \quad . \quad (14)$$

The relationship between mean-square heater current and conducting angle is shown graphically in Fig. 10, and applies for either heater if the appropriate suffix is attached.

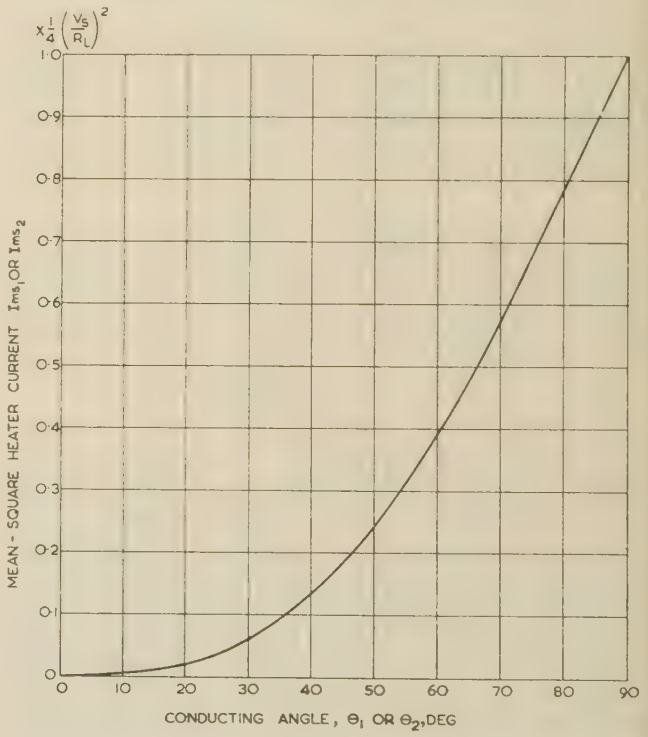


Fig. 10.—Mean-square heater current versus conducting angle.

The heater currents of thermistors TH_1 and TH_2 pass alternately through the heater of TH_3, and the latter is shunted to make

$$I_{ms3} = \tfrac{1}{2}(I_{ms1} + I_{ms2}) \quad . \quad . \quad . \quad . \quad . \quad (15)$$

which remains true under all conditions. The dissipations in the heaters of thermistors TH_1 and TH_2 are thus increased and decreased, respectively, by equal amounts by the a.c. drive, being equal to that in the heater of TH_3 when the drive is zero.

The overall d.c. feedback is arranged so that, when the ambient temperature is midway between its upper and lower limits, i.e. at 30° C in this case, I_{ms3} is held equal to one-half of the maximum possible value of I_{ms1} or I_{ms2}, i.e.

$$I_{ms3} = \frac{1}{8}\left(\frac{V_s}{R_L}\right)^2 \text{ at } 30^\circ\text{C} \quad . \quad . \quad . \quad . \quad (16)$$

Combining this with eqn. (15),

$$I_{ms1} + I_{ms2} = \frac{1}{4}\left(\frac{V_s}{R_L}\right)^2 \text{ at } 30^\circ\text{C} \quad . \quad . \quad . \quad (17)$$

In practice, all three heaters receive 15 mW when the ambient temperature is 30° C and there is no drive. When drive is applied, the dissipation in one of the output thermistors rises towards a limit of 30 mW, while that in the other falls towards zero.

Combining eqns. (13), (14) and (17) we obtain the following

expression, which interrelates the two conducting angles when operation is at the design-centre ambient temperature of 30° C:

$$\frac{2\theta_1 - \sin 2\theta_1}{\pi} + \frac{2\theta_2 - \sin 2\theta_2}{\pi} = 1 \quad . \quad . \quad . \quad (18)$$

This relation is shown graphically in Fig. 11. Eqn. (12) now permits I_s/I_b to be related to conducting angle as shown in Fig. 12.

Since the mean-square heater current is known in terms of conducting angle, it can be related to I_s/I_b, as shown in Fig. 13. This gives the heater dissipations in terms of the signal, whence the bead resistances and the voltage divider percentage output can be calculated for any particular thermistor characteristics. Fig. 6 shows the result for thermistors whose resistance/dissipation characteristic is given in Fig. 14, where $V_s = 10\sqrt{2}$ volts,

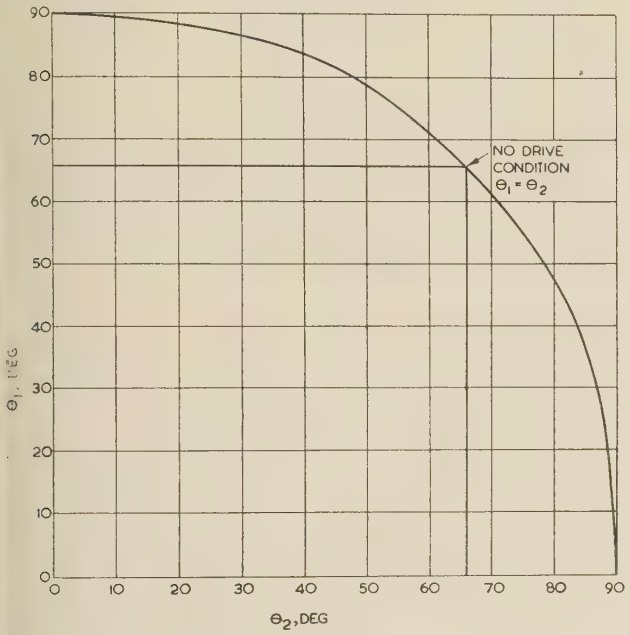


Fig. 11.—Relation between conducting angles at 30° C.

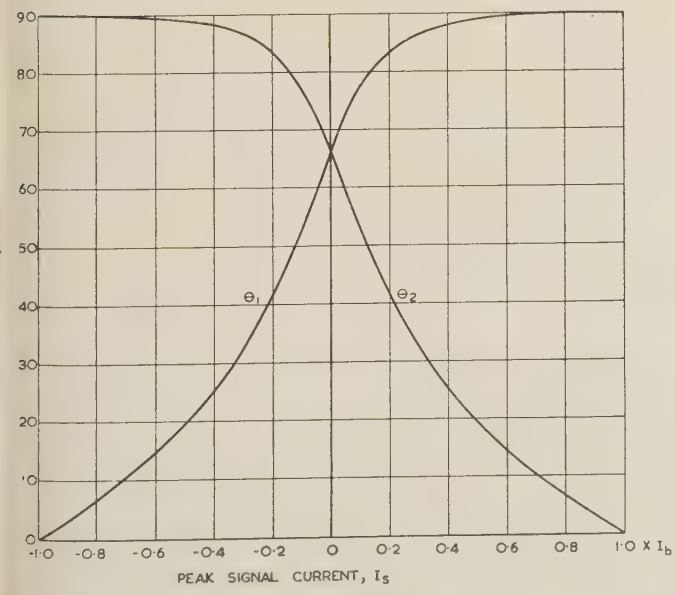


Fig. 12.—Conducting angle versus signal at 30° C.

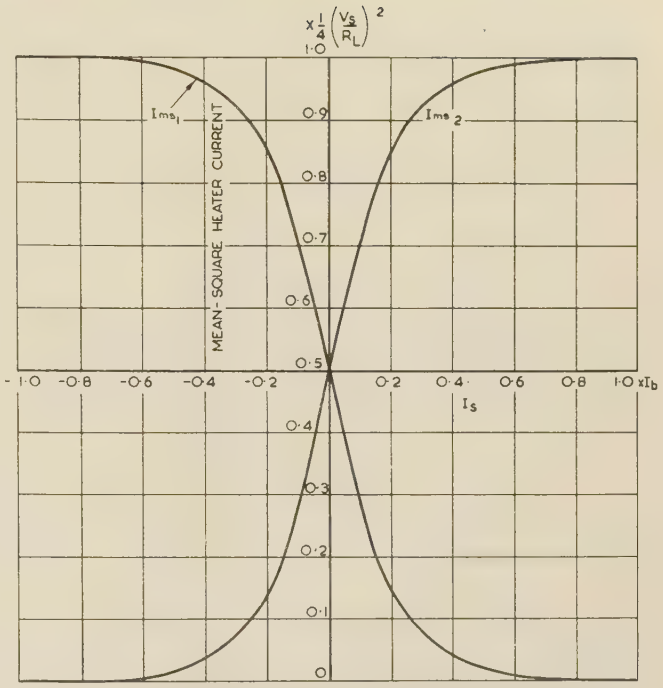


Fig. 13.—Mean-square heater current versus signal at 30° C.

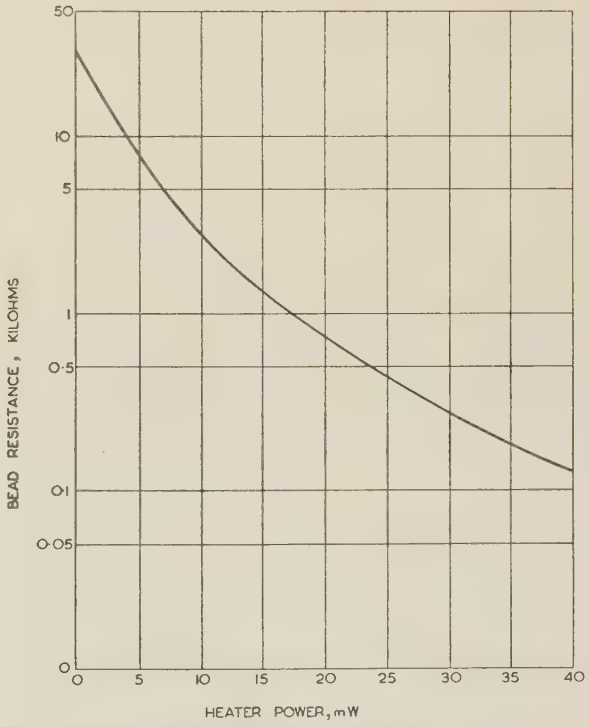


Fig. 14.—Characteristic of thermistor at 30° C.

$R_L = 410$ ohms, $\frac{1}{8}(V_s/R_L)^2 = 149(\text{mA})^2$, and each heater (since its resistance is 100 ohms) dissipates 15 mW at 30° C when there is no a.c. drive.

The demodulator characteristic shown in Fig. 6 is strictly true for operation at 30° C only, but the characteristics at 0° C and 60° C are not significantly different within the limits of ±75% of the theoretical maximum output.

As can be seen, the characteristic is reasonably linear up to $\pm 75\%$ of full output, at which point $I_s/I_b = 0{\cdot}1$, approximately. Since I_b is in practice equal to 0·5 mA, the value of I_s corresponding to 75% output is 0·05 mA, and the r.m.s. value of the a.c. signal applied to the base of J_3 is $I_s/\sqrt{2} = 0{\cdot}032$ mA, as was stated in Section 4.2.

(7.3) Calculation of TH_3 Shunt Resistance

It is required that

$$I_{ms3} = \tfrac{1}{2}(I_{ms1} + I_{ms2}) \quad . \quad . \quad . \quad . \quad . \quad (19)$$

Let the heater of thermistor TH_3 be shunted by a resistance R, to a value R_s. Since the heaters of thermistors TH_1 and TH_2 receive current during alternate half-cycles, the mean-square collector current of J_4 is equal to $I_{ms1} + I_{ms2}$. The mean-square current in the heater of TH_3, whose resistance is 100 ohms, is thus

$$I_{ms3} = (I_{ms1} + I_{ms2})\left(\frac{R_s}{100}\right)^2 \quad . \quad . \quad . \quad . \quad (20)$$

Combining eqns. (19) and (20), $R_s = 100/\sqrt{2} = 70{\cdot}7$ ohms, whence the shunt resistance, R, is 242 ohms.

(7.4) Inter-Stage Bias Arrangements

Inter-stage biasing is arranged to ensure that transistors J_3 and J_4 can be switched on or off at all parts of the cycle by currents applied to J_1, in ambient temperatures up to at least 60° C.

A positive d.c. bias of 0·6 mA is supplied to the base of J_4, ensuring that it can be switched off by J_3 even though its collector leakage rises to 0·6 mA. This allows a junction temperature of 75° C if J_4 is a type OC76 transistor, and an ambient temperature of about 70° C is permissible since the dissipation in the transistor is low.

R_{14} limits the collector current of J_3 to about 2 mA, which is ample to switch J_4 on and cannot overload J_3. A positive bias of 0·6 mA is also supplied to the base of J_3. In this case, about 0·2 mA may flow through R_{11} to provide adequate cut-off voltage for J_3 when J_2 is bottomed. Thus the J_3 leakage must not exceed 0·4 mA and its junction temperature should not exceed 70° C. J_3 dissipation is very low, however, and allows an ambient temperature of nearly 70° C.

Even if J_2 collector leakage is 0·4 mA, rather more than 1·3 mA is drawn from the base of J_3 when J_2 is switched off. J_3 can therefore be switched on throughout the cycle, since the sum of the positive d.c. bias and the peak value of the rectified bias is less than 1·3 mA.

The emitter resistor of J_2 carries a minimum of 0·7 mA, which provides adequate collector voltage for J_1 and ensures that J_2 can be switched on and off. The dissipation in J_1 and J_2 is very low, and both can be operated at 70° C provided that J_1 then has a collector leakage (below about 30 μA) which can be supplied by the overall d.c. feedback. Most transistors of the OC45 type meet this requirement, and the few whose leakage is close to the specified upper limit meet it at 60° C.

DISCUSSION ON 'A SURVEY OF PERFORMANCE CRITERIA AND DESIGN CONSIDERATIONS FOR HIGH-QUALITY MONITORING LOUDSPEAKERS'*

NORTH MIDLAND CENTRE, AT LEEDS, 3RD FEBRUARY, 1959

Mr. E. M. Price: It is gratifying that the author comes out firmly in favour of high-grade loudspeakers for monitoring purposes.

In sound reproduction it is amazing that the early pioneers with their simple equipment achieved intelligibility at all, while the effort put into the loudspeaker problem since that time has achieved, in general, neither finality in design nor near perfection in performance. Many fine performances have been demonstrated, but they have been in the nature of special efforts.

In Section 2.3 it is suggested that the sound reaching the ear from different directions produces stimuli which are integrated without conscious effort by the listener. This seems to imply that the air-pressure variation at the listener's ear is not a complete objective description of the sound at that point, even for monaural listening. If this is so, it seems that more research is necessary on the manner in which the ear receives and interprets sound waves.

Similarly, ignorance of the mode of operation of the ear affects the consideration of phase distortion. The suggestion that, in some instances, the loudspeaker behaves like a non-minimum phase device is interesting, but what inferences are to be drawn from this? It may result in more phase distortion, but this does not alter the nature of the problem. It seems again that more knowledge of the behaviour of the ear to transient sounds is required, and it may be that further substantial progress depends on combined physiological and acoustical research.

Can the author give more information on noise-band testing? It sounds attractive, but what does the listener observe as a result of differences between loudspeakers?

Mr. H. D. Kitchin: The author discusses the importance of the directional characteristic of the loudspeaker system, particularly at the higher audio frequencies, but in relating this to the directional characteristic of the original source he only mentions some conclusions reached by Kaufmann. However, one would expect different instruments and instrumental combinations to differ considerably in their directional properties. For example, horn instruments should be more directional than stringed instruments. Presumably, corresponding directional characteristics are necessary in the reproducing system if one is to achieve the utmost realism. As it would seem impossible to devise a single loudspeaker system which will change its directional characteristic to suit the instrumental combination being reproduced, the results must always be a compromise. I wonder whether the author has more information on this point from his own investigations.

In Section 2.7.3 it is stated that speech picked up by a microphone at close quarters acquires an unnatural character which cannot be remedied by electrical equalization. This effect is therefore not entirely due to frequency discrimination. Can the author suggest the nature of this distortion?

All high-fidelity enthusiasts must envy the opportunity presented at broadcasting and recording companies to compare the reproduced sound with the original source, but, in the sphere of popular music, such comparisons are likely to be of little value, because of the many acoustic and electronic tricks often employed. It seems rather difficult to define precisely what is meant by 'high fidelity', as it is certainly not fidelity to the original sound as performed by the artists. Can the author suggest how we may appraise the quality of reproduction in these cases?

Mr. A. J. Coveney: Would the author give us his views on the performance of the 'column' type of loudspeaker? The units now available seem to have a full 360° sound transmitting angle on the horizontal axis, but only a small angle on the vertical axis. Therefore, in some instances, they would be preferable to the orthodox type of box-cabinet speaker.

If correctly designed, they appear to have the effect of reducing 'boom' and over-accentuating the base notes, which, with orthodox loudspeakers, sometimes tend to drown the treble notes and those on the higher-frequency end of the scale. The overall effect is that they give a clearer reproduction of the original sounds.

Mr. D. E. L. Shorter (*in reply*): In reply to Mr. Price, all the remarks in the paper about the subjective summation of direct and indirect sound relate to binaural listening. Here the sense of direction undoubtedly plays a large part so that the brain receives information over and above that obtainable from a pressure measurement at one point.

The question whether a given loudspeaker does or does not act as a minimum phase-shift device must be considered in the context of the age-old controversy as to whether or not the steady-state frequency characteristic should be taken as the sole criterion of the response; the practical effect of the answer is to decide whether some other kind of measurement is required if we are to account for what we hear. It is debatable whether the next step towards closing the gap in our knowledge should be an extension in the scope of our measurements of loudspeaker performance or further research on the mechanism of hearing. Ultimately we shall need both, but it would not be unfair to say that if the development of electro-acoustic devices had been dependent on hearing research the art would be even more backward than it is.

Subjective testing with noise is usually carried out, not with selected frequency bands, but with the complete frequency range presented at once. The noise represents a 'programme', the original of which is not accessible for comparison, but this limitation is not as serious as might be supposed, since the test is very sensitive and even good loudspeakers will be found to give recognizable emphasis to one or more regions of the frequency band. The technique is most useful for comparative purposes, but an experienced observer can diagnose faults on a single loudspeaker on the basis of his past experience.

I share Mr. Kitchin's concern at the meagre amount of published information on the directional characteristics of loudspeakers. Because musical instruments differ in their directional properties we cannot hope to produce a rigorously 'correct' impression of a performance in the listening room (see Section 2.1). However, in orchestral music, the polar characteristics of individual instruments have less direct effect than might be supposed because a high proportion of the sound heard in the studio is

* SHORTER, D. E. L.: Paper No. 2604 R, April, 1958 (see 105 B, p. 607).

reflected from the walls, floor and roof. The effect of a widely spread source such as an orchestra can be, to some extent, simulated by the omnidirectional type of loudspeaker described in the paper and, to a lesser degree, by the column loudspeakers referred to by Mr. Coveney. However, the effect depends very much on the characteristics of the room, and it is, in any case, inappropriate for the reproduction of speech. Since the human voice is here, as in other respects, the chief stumbling block in the pursuit of realism, it might be well to take the directional characteristics of speech as a basis for compromise in the design of loudspeakers, if this could be done while satisfying the other requirements.

The peculiarity in voice quality at close range referred to in Section 2.7.3 arises because the mouth behaves as a point source whose position varies according to the frequency and the particular speech sound being pronounced; the situation is further complicated by the variable contribution made by the sound emanating from the nostrils as well as that directly radiated from the chest. Any equalization of the frequency characteristic can therefore be only approximate, and the resulting quality, while acceptable for many purposes for which close-talking microphones have to be used, is not adequate as a standard for the testing of loudspeakers.

Where, as mentioned in Section 2.7.4, special microphone arrangements are used to produce effects which cannot be heard in the studio, there is, of course, no basis for an absolute subjective assessment of loudspeaker quality. However, if these arbitrary effects are to be reproduced as originally intended, some degree of uniformity in the response of loudspeakers must be aimed at. By choosing as standard those characteristics which give the best results when comparison with the original *is* possible, both sets of requirements are simultaneously satisfied.

DISCUSSION ON
'A NEW CATHODE-RAY TUBE FOR MONOCHROME AND COLOUR TELEVISION'*

SHEFFIELD SUB-CENTRE, AT SHEFFIELD, 17TH DECEMBER, 1958

Prof. A. L. Cullen: In what respect is the use of a Permalloy plate dividing the two halves of the tube an advantage? Why is a magnetic screen necessary?

In Fig. 5 it is shown that the beam is brought to a focus before reaching the screen in order to have a collimated electron beam impinging on the shadow mask. However, this collimation is only effective in the Y-direction. Presumably there is a defocusing action in the X-direction. Is this of any importance?

In Section 4 it is stated that the 3-colour gun has three cathodes side by side, with a common heater, and that this facilitates the use of impregnated high-density cathodes. I do not understand the reason for this.

In Section 8.3 a gettering area of $2\frac{1}{2}$ ft^2 is mentioned. How do the authors envisage this area being used effectively?

I would like to have information about the accuracy of the photo-etching technique used in constructing the shadow mask. Clearly the dimensions involved are extremely small, and it would be interesting to know the accuracy to which they can be maintained.

The field flyback depends for its operation on secondary emission. Have the authors found any variability in the secondary-emission coefficient owing to contamination of the electrodes during the periods when the system is opened to the air? In experiments with Dr. I. M. Stephenson on a klystron valve, we were troubled with secondary emission, and found that it was dependent upon the surface condition of the metal, as I think is well known. In our case oil pumps were used, and certainly oil vapour was the major cause of the contamination.

I found the design of the reversing lens on geodesic principles fascinating. A parallel in the handling of electromagnetic waves rather than electron beams is given by the Schwarzchild rapid-scanning antenna† developed during the war.

During a visit to Prof. Gabor's laboratory at an early stage in the development of the tube, I was greatly impressed by the very convenient vacuum tanks which were used for the development work. I imagine that it would have been almost impossible to carry out so many electrical optical experiments in such a relatively short time without them. It would be of interest to have full details of the design of these vacuum tanks, together with some information on the authors' experience in using them.

Prof. D. Gabor and **Drs. P. R. Stuart** and **P. G. Kalman** (*in reply*): A magnetic screen is not essential in the monochrome tube, but in the colour tube it is important to suppress components of the magnetic field parallel to the screen's surface, as this might produce spurious colours.

Collimation of the electron beams in the X-direction is unimportant, because the slits in the shadow mask are horizontal.

In order to obtain good horizontal focusing over the whole of the picture area it is necessary to compensate the X-defocusing of the collimator and reversing lens by the introduction of the correct amount of X-focusing from the line-scan deflectors. X-deflectors of suitable design are shown in Fig. 11(*c*) of the paper.

Impregnated cathodes suffer from the disadvantage that they require rather high temperatures, and there is an appreciable saving in heating power if all cathodes are served by a common heater.

With sheets of 0·001 in thickness it is not difficult to maintain in photo-etching an accuracy of 0·000 5 in.

There are fortunately some secondary-emitter materials which are not sensitive to contamination by contact with atmospheric air. Silver sulphide, which we used in our first experiments, is one of these, while others are beryllium copper and beryllium nickel.

Shortage of space makes it impossible to give details of the vacuum tanks. Our own system, although convenient for the purpose, is now far superseded by those at the A.E.R.E., Harwell.

* GABOR, D., STUART, P. R., and KALMAN, P. G.: Paper No. 2661 R, May, 1958 (see 105 B, p. 581).

† RIDENOUR, L. N.: 'Radar System Engineering'. Radiation Laboratory Series Vol. I (McGraw-Hill), p. 295.

PROCEEDINGS OF THE INSTITUTION OF ELECTRICAL ENGINEERS

Part B. ELECTRONIC AND COMMUNICATION ENGINEERING (INCLUDING RADIO ENGINEERING), JANUARY 1960

CONTENTS

	PAGE
The President's Inaugural Address SIR WILLIS JACKSON, F.R.S.	1
Centre Chairmen's Addresses	14
Discussion on 'A Train Performance Computer'	19
Discussion on 'A Train Performance Computer'	20
The Performance of a Balanced Aerial when connected directly to a Coaxial Cable G. D. MONTEATH, B.Sc., and P. KNIGHT, B.A.	21
Classes of 4-Pole Networks having Non-Linear Transfer Characteristics but Linear Iterative Impedances. PROF. E. COLIN CHERRY, D.Sc.(Eng.)	26
Phase-Shift Keying in Fading Channels H. B. VOELCKER	31
An Analysis of a Type of Comb Filter A. G. J. MACFARLANE, B.Sc.	39
The Measurement of Atmospheric Radio Noise by an Aural Comparison Method in the Range 15–500 kc/s. J. HARWOOD, M.A., Ph.D., and B. N. HARDEN, M.Sc.	53
An Experimental Investigation of a Two-Cavity Klystron operating under Large-Signal Conditions. I. M. STEPHENSON, Ph.D., M.Sc., B.Sc.	60
The Use of a Wire-Wound Helix to form a Circular H_{01} Wavemeter Cavity. PROF. H. E. M. BARLOW, Ph.D., B.Sc.(Eng.), H. G. EFFEMEY and P. H. HARGRAVE, B.Sc.	66
A Coaxial Film Bolometer for the Measurement of Power in the U.H.F. Band I. A. HARRIS	67
A Transistor Quadrature Suppressor for A.C. Servo Systems I. C. HUTCHEON, M.A., and D. N. HARRISON	73
Discussion on 'A Survey of Performance Criteria and Design Considerations for High-Quality Monitoring Loudspeakers'	83
Discussion on 'A New Cathode-Ray Tube for Monochrome and Colour Television'	84

Declaration on Fair Copying.—Within the terms of the Royal Society's Declaration on Fair Copying, to which The Institution subscribes, material may be copied from issues of the *Proceedings* (prior to 1949, the *Journal*) *which are out of print and from which reprints are not available.* The terms of the Declaration and particulars of a Photoprint Service afforded by the Science Museum Library, London, are published in the *Journal* from time to time.

Bibliographical References.—It is requested that bibliographical reference to an Institution paper should always include the serial number of the paper and the month and year of publication, which will be found at the top right-hand corner of the first page of the paper. This information should precede the reference to the Volume and Part.

Example.—SMITH, J.: 'Reflections from the Ionosphere', *Proceedings I.E.E.*, Paper No. 4001 R, December, 1954 (102 B, p. 1234).

There are many members and former members of The Institution who are finding life difficult. **When remitting your membership subscription,** please help them by sending a donation, or annual subscription preferably under deed of covenant, to

THE BENEVOLENT FUND

The object of the Fund is to afford assistance to necessitous members and former members (of any class) of The Institution of Electrical Engineers who have paid their subscriptions for at least five years consecutively or compounded therefor, and to the dependants of such members or former members.

Subscriptions and Donations may be sent by post to

THE INCORPORATED BENEVOLENT FUND OF THE INSTITUTION OF ELECTRICAL ENGINEERS, SAVOY PLACE, LONDON, W.C.2

or may be handed to one of the Local Hon. Treasurers of the Fund.

Though your gift may be small, please do not hesitate to send it

Local Hon. Treasurers of the Fund:

EAST MIDLAND CENTRE *H. J. McLean*	SCOTTISH CENTRE *R. H. Dean, B.Sc.Tech.*
IRISH BRANCH *A. Harkin, M.E.*	NORTH SCOTLAND SUB-CENTRE *P. Philip*
MERSEY AND NORTH WALES CENTRE *D. A. Picken*	SOUTH MIDLAND CENTRE . . . *H. M. Fricke*
TEES-SIDE SUB-CENTRE *W. K. Harrison*	RUGBY SUB-CENTRE *P. G. Ross, B.Sc.*
NORTH-EASTERN CENTRE . . . *J. F. Skipsey, B.Sc.*	SOUTHERN CENTRE *G. D. Arden*
NORTH MIDLAND CENTRE *E. C. Walton, Ph.D., B.Eng.*	WESTERN CENTRE (BRISTOL) . . . *A. H. McQueen*
SHEFFIELD SUB-CENTRE *F. Seddon*	WESTERN CENTRE (CARDIFF) . . . *E. W. S. Watt*
NORTH-WESTERN CENTRE . *E. G. Taylor, B.Sc.(Eng.)*	WEST WALES (SWANSEA) SUB-CENTRE *O. J. Mayo*
NORTH LANCASHIRE SUB-CENTRE *H. Charnley*	SOUTH WESTERN SUB-CENTRE . . *W. E. Johnson*
NORTHERN IRELAND CENTRE . . *G. H. Moir, J.P.*	

THE BENEVOLENT FUND

Members are asked to bring to the notice of the Court of Governors any deserving cases of which they may have knowledge.

Published by The Institution, Savoy Place, London, W.C.2. *Telephone:* COVent Garden 1871. *Telegrams:* 'Voltampere, Phone, London.'
Printed by Unwin Brothers Limited, Woking and London.